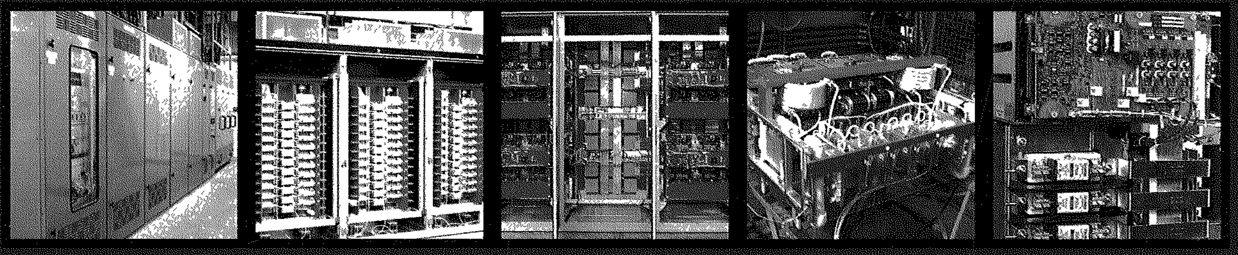
Medium Voltage Variable Frequency Drives For Induction and Synchronous Motors



TECHNICAL PAPER

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Introduction

MEDIUM VOLTAGE VARIABLE FREQUENCY DRIVES FOR INDUCTION AND SYNCHRONOUS MOTORS

The global power electronics industry continues the rapid pace of solid-state drive development. Over the past four decades, many drive circuits have become virtually obsolete and new ones introduced. The user is confronted with a wide variety of drive types that are suitable for virtually every kind of electrical machine from the sub-fractional to the multi-megawatt rating. In this paper we will concentrate on commercially available drives suitable for operating a standard medium voltage polyphase AC motor.

This category constitutes a major consumer of electric power in industrial applications, and represents the opportunity for substantial improvement in the user's process, as well as energy savings. Both new installations and the retrofit of existing machines are possible.

Despite the diversity of power circuits, these drives have two common properties:

1. All accept commonly available AC utility power of fixed voltage and frequency, and through switching power conversion, create an output of suitable characteristics to operate a particular type of electric machine—in this case 3-phase AC.

2. All are based on solid-state switching devices. The development of new devices drives this technology. This paper will illustrate the characteristics of commonly used devices.

Figure 1 illustrates the basic structure of most common AC drives. The input conversion circuit converts the utility power, which has a constant frequency and amplitude, into DC. An output inversion stage changes the DC back into AC with variable frequency and amplitude. Other elements shown in the diagram are optional.

There are a number of reasons to use a variable speed drive:

1. Energy savings where variable flow control is required. Any situation in which flow is controlled by a throttling device (valve or damper) has the potential for energy savings by removing the throttle and slowing the fan or pump to regulate flow.
2. Optimizing the performance of rotating equipment; *e.g.,* SAG mills, compressors, conveyors, pumps and fans.
3. Elimination of belts and gears or other power transmission devices by matching the base speed of the motor to the driven load.
4. Automation of process control by using the VFD as the final control element—leading to more efficient part-load operation.
5. Reduction of the rating and cost of the electrical distribution system by eliminating motor starting inrush.
6. Extending the life of motors, bearings, seals, liners and belts.
7. Reducing noise and environmental impact. Electric drives are clean, non-polluting, quiet, efficient and easy to repair.

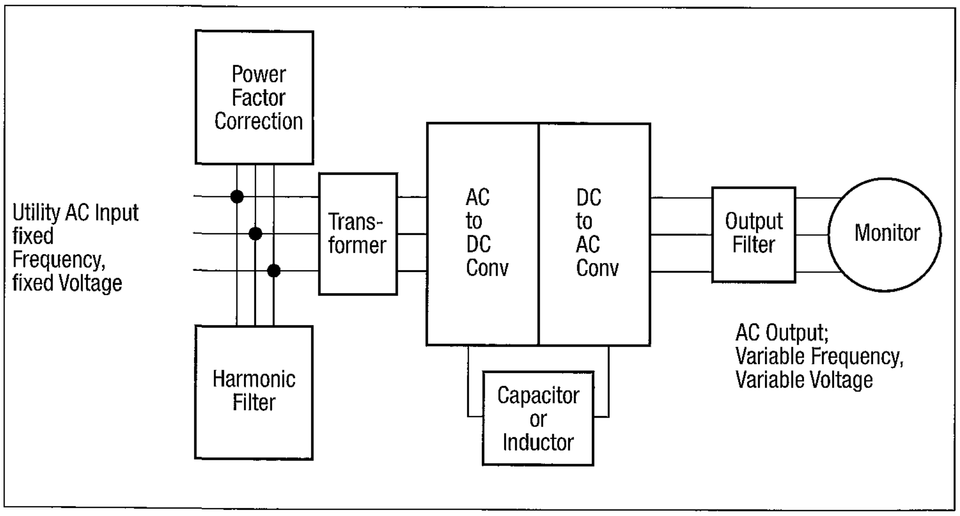

Figure 1. Structure of a generic variable frequency drive

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Semiconductor Switching Devices

Even though many of the basic power conversion principles were developed in the 1930s, when circuits were constructed with mercury arc rectifiers, it was not until the invention of the thyristor in 1957 that variable speed drives became truly practical. The semiconductor devices discussed in this paper are constructed of silicon with two to four layers and different doping. Silicon devices are presently limited to less than 10kV blocking voltage and have a maximum operating temperature of 150C, although some are as low as 100C.

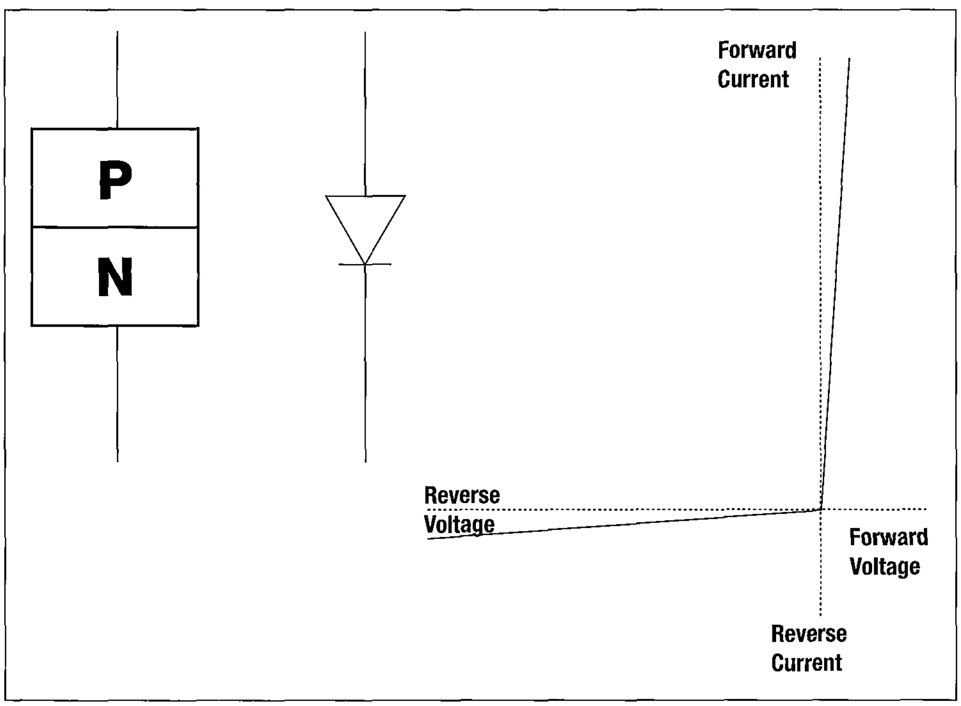

Figure 2. Structure and properties of the diode or rectifier

The most elementary silicon diode or rectifier is shown in Figure 2. There are two "layers" of silicon and one junction. ("Layers" defines a convenient way to visualize the device, but it is actually one monolithic crystal of silicon with layers of different doping—hence conductivity.) Because of the doping, the diode behaves as a one-way device, allowing current to flow freely in one direction, but blocking in the opposite. Typically, the silicon diode has a voltage drop when conducting 0.5v to 1.5v, depending on the voltage rating. Higher voltage-rated units have higher forward drop because they are constructed of thicker silicon. The current increases exponentially with forward voltage, resulting in less change in voltage over a large range of current. When blocking reverse voltage, the leakage current is quite small by power standards, generally much less than 50mA. However, the leakage current of a number of diodes are not likely to be close to the same value; this is also true for other types of semiconductors. The properties of silicon devices, such as forward drop and leakage current, are quite sensitive to temperature. The diode has no control input, so it is considered a passive device, and a non-linear one.

In order to control the circuit, active devices are required—with a control input. By applying a low power signal we can make these devices turn on or off. Ideally, a power switch would have zero on-state voltage and zero off-state leakage current. It would be capable of changing state instantaneously and could carry current or block voltage in either direction. Although much progress has been made, power semiconductors do not yet approach this ideal behavior.

Fortunately, practical circuits utilizing available device properties exist. The voltage-fed inverter requires devices that conduct in both directions, but need only block voltage in one direction. The current-fed inverter needs switches that can block voltage in both directions, but conduct current in only one direction.

The thyristor (SCR) is a four-layer semiconductor device that has some of the properties of an ideal switch. It has low leakage current (at most 10s of mA) in the off-state, a small voltage drop in the on-state (1 to 3 V), and takes only a small current signal to initiate conduction. Power gains of over 10^6 are common. When applied properly, the

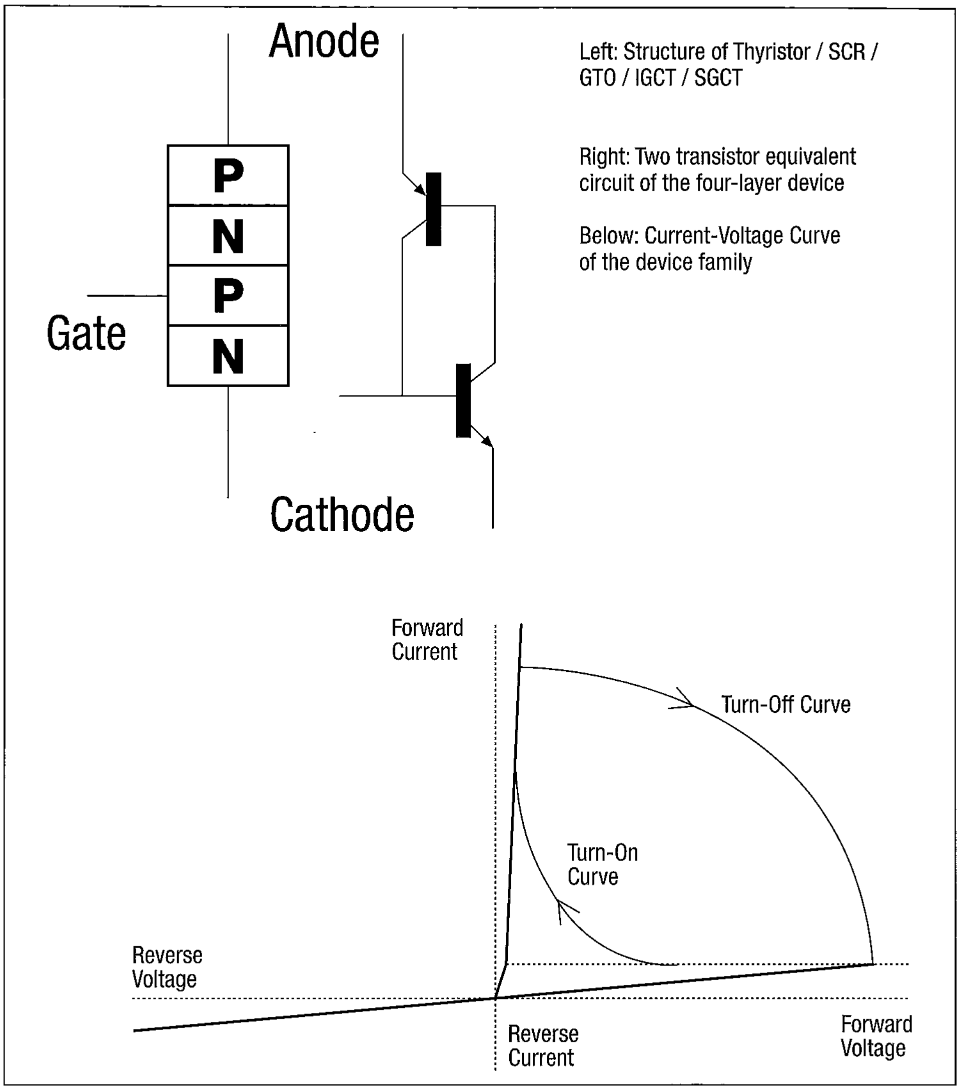

Figure 3. Properties of 4-layer devices such as the thyristor, gate-turn-off thyristor. IGCT, and SGCT

thyristor will last indefinitely. After its introduction, the current and voltage ratings increased rapidly. Today it has substantially higher power capability than any other solid-state device, but no longer dominates power conversion in the medium and higher power ranges. The major drawback of the thyristor is that it cannot be turned off by a gate signal. The anode current must be interrupted in order for it to regain the blocking state. The inconvenience of having to commutate the thyristor in its anode circuit at a very high energy level has encouraged the development of other closely related devices as power switches.

The thyristor or SCR and its siblings the IGCT and SGCT do not have a linear region. They are intended solely as a two-mode switch—either on or off. Once conduction is initiated, the internal feedback mechanism maintains the On state even if the gate current is shut Off. This characteristic also means that the turn On is extremely fast and not controllable. However, note the V-I curve in Figure 3. It cannot conduct current in the reverse direction, even though it can block voltage in both directions. The first generation of variable frequency drives were made with thyristors. Various strategies were used to turn Off the devices. Usually, this requires stored energy in a capacitor to be discharged by another thyristor. The thyristor can be regarded as a mature technology manufactured by dozens of companies.

It has been known, almost from the invention of the thyristor, that some types could be turned Off by using a relatively large negative gate current. This interrupts the internal positive feedback that keeps the device in the On state. A class of devices, known as Gate Turn Off Thyristors or GTOs, was developed into practical products. Because of thyristor structure, GTOs can have blocking voltages as high as 6kV and can turn off as much as several thousand amperes of current. The turn-off gain is around 3 to 4; that is, it takes 1/4 to 1/3 of the anode current as a negative current in the gate circuit to achieve turn-off. Even though this can be a large current, the negative gate voltage is around 20v. So the gate power is very much less than the main power. But the capability to turn-off came with a price. The on-state voltage drop increased substantially and fabrication become much more difficult. Furthermore, it was discovered that the turn-off process causes large heat losses in the silicon. Each turn-on or turn-off causes an irretrievable energy loss in the silicon. This is true for all semiconductor switches (see Figure 8). In the case of the GTO, the switching losses limited the maximum switching frequency to 200 – 300 Hz. In order to achieve desired performance, GTOs were most often used with large R-C snubbers which moved the losses from the silicon to external components. Since a large proportion of GTOs were used in voltage-fed inverters, the reverse-blocking capability was sacrificed to improve other characteristics. In order to achieve desired waveforms, rapid switching is very advantageous. This requirement was inconsistent with large switching losses in the GTO. Nevertheless, many variable frequency drives based on GTOs built in the late '80s had good success. However, the high cost and very large switching losses partially restricted the use of GTOs to only those applications in which space and weight were at a premium.

Because of fabrication difficulty and the relatively small demand, power GTOs were manufactured by fewer than a half-dozen companies—and the number of suppliers is shrinking.

More recently, the use of IGCT, an improved means of turning off GTOs, was introduced. The principle of "hard drive" goes back to about 1980, but was only introduced in high-power devices in the past five years. By using a very fast and powerful gate driver located on top of the device, the anode current can be drawn out quickly (1us) through the gate. This prevents much of the device loss encountered in the GTO turn-off. The turn-off gains unity. The turn-off pulse must be very fast-rising to minimize device losses. And the driver must be close to reduce the gate-

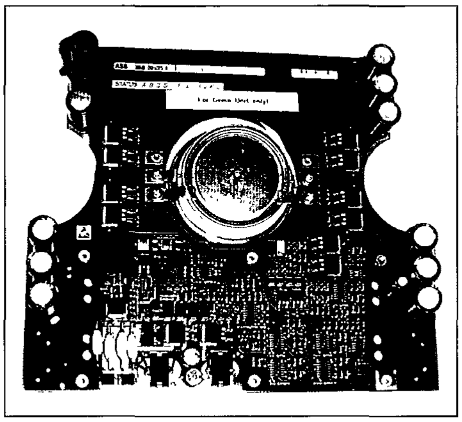

Figure 4. An IGCT showing the puck housing for the device itself and the surrounding circuit, which is the integrated gate driver. There are eight MOSFETs around the device that switch the gate current. Nine electrolytic capacitors at the edge of the board store energy for the gate pulse.

circuit inductance. Having no linear region, the IGCT is a member of the thyristor family and is therefore suitable only for switching. Like the thyristor, the IGCT has internal positive feedback. Once triggered it will remain On until turned Off, either by a large negative gate pulse or by the reduction of the anode current to zero. Figure 4 shows an IGCT with the gate driver circuit surrounding the device.

A further subtlety of the GTO/IGCT family is the reverse blocking capability. In voltage-fed circuits the switches do not need to support reverse voltage, but current-fed circuits require devices with reverse blocking. Since the device design is a trade-off among forward drop, blocking voltage, turn-off capability and speed, both GTO and IGCTs are available in asymmetrical versions, which have no reverse blocking ability. This allows for better optimization of other characteristics. For the current-fed circuits, there are symmetrical IGCTs, which have both forward and reverse blocking. These are known as SGCTs (Symmetrical Gate Controlled Thyristor).

Transistors pre-date thyristors, but their use as high-power switches was relatively restricted (compared with thyristors) until the ratings reached 50 A and 1,000 V in the same device, during the early 1980s. Bipolar transistors are three-layer semiconductors that exhibit linear behavior but are used only in saturation (fully turned on) or fully turned off. The transistor is turned on by a base current, which must be maintained to keep the device in conduction. In order to reduce the base drive requirements, most transistors that were used in variable speed drives are Darlington types, which have a pre-amplifying transistor ahead of the main one. Even so, they have higher conduction losses and greater drive power requirements than thyristors. Nevertheless, because they can be turned on or off quickly via base signals, transistors quickly displaced thyristors in lower drive ratings, and were once widely used in pulse-width modulated voltage source inverters. Figure 5 shows the structure and response curve of a bipolar transistor. They in turn were displaced by insulated gate bipolar transistors (IGBTs) in the late 1980s. The IGBT is a combination of a power bipolar transistor and a MOSFET (see Figure 6) that combines the best properties of both devices. A most attractive feature is the very high input impedance that permits them to be driven directly from lower power logic sources. The MOSFET (metal-oxide-semiconductor-field-effect-transistor) can be thought of as the driver transistor. As voltage on the gate increases, current flows through the MOSFET and into the base of the complementary PNP bipolar transistor. The device is normally off and it can operate as a linear amplifier. This capability has been useful in controlling di/dt in circuits and sharing voltage in series strings of IGBTs. In addition to the high-input impedance, the IGBT does not have as much stored charge as the IGCT or transistor, and is therefore a significantly faster switching device. The on voltage drop is somewhat higher than that of an IGCT of the same voltage rating. The IGBT and bipolar transistor have no reverse voltage blocking ability, so they are most

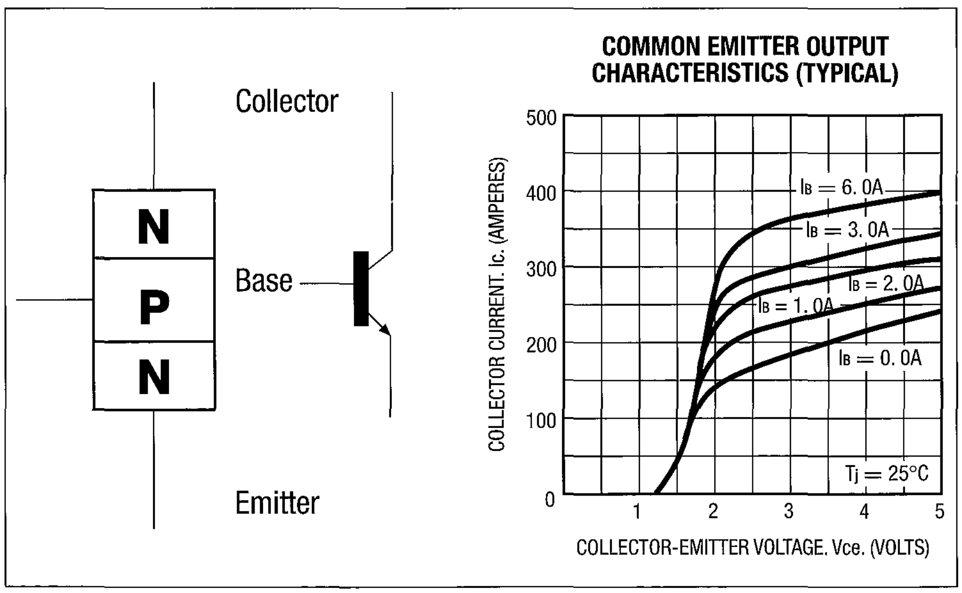

Figure 5. Structure and response of a 300A 1200V bipolar power transistor

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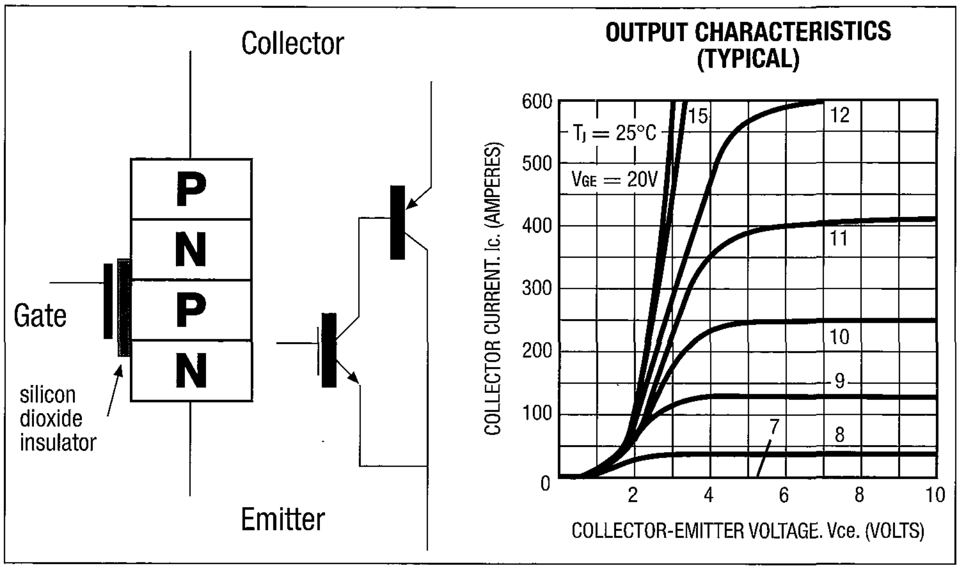

Figure 6. Structure and response of an IGBT

often used in voltage source circuits with a diode connected in reverse parallel. Unlike the other devices, power IGBTs are actually arrays of thousands of tiny devices in parallel.

The power handling capability of IGBTs has increased dramatically in the past 5 years. Presently, 3300v devices at 1200A are widely available with higher ratings on the way. The forward voltage drop has declined steadily as the result of newer processing technology and geometry such as the trench-gate.

IGBTs are now viable alternatives to thyristors, GTOs and IGCTs in the largest drive ratings. There are more than a dozen manufacturers of power IGBTs and many improvements in device characteristics have been introduced. Because of the huge numbers of IGBTs used in low-voltage VFDs, they enjoy the cost and reliability benefits of mass production.

Very recently another device has emerged, which is known as the IEGT (injection-enhanced gate transistor). This is a specially constructed IGBT which has a low on state voltage and is presently capable of blocking up to 4.5kV. It appears to be aimed at deployment in the neutral-point-clamped inverter for medium voltage inverters. It is manufactured by only one Japanese company.

How is relative performance of semiconductor switches compared? One obvious way is to compute the product of the rated voltage and rated current which gives a rough idea of the VA capacity of the device. But there are other factors to consider. The device switching speed is important not only because it permits better waveforms, but it is also strongly (inversely) related to switching losses. The faster a device can traverse through the region between on and off, the lower the switching losses. Figure 7 shows the turn-off switching event; the current decreases as the voltage across the device increases. In the middle of the event, there is simultaneously high current and voltage, which represents power being dissipated in the silicon. In an application, one has to look at both the conduction losses due to the on-state voltage drop and also the switching losses. Certain categories of device, *e.g.*, the IGCT and the IGBT, are often characterized by the maximum turn-on and turn-off energy at specified operating conditions. Then the switching loss is the product of the energy per switching event times the operating frequency. The total device losses are very important in determining the size, weight and efficiency of the circuit.

Another factor is maximum operating temperature at which the performance is guaranteed. This is usually 150C for diodes and 125C for IGBTs and thyristors. IGCTs are rated at slightly lower temperature.

In the early days of power semiconductors, the power semiconductors were housed in ceramic packages which had a threaded stud on the bottom to attach the device to its heatsink. Since all the switches have power losses and a maximum operating temperature, it is almost always necessary to provide a "heat sink" to carry away the power loss and maintain a suitable working temperature. The heat is carried away by air or by water. A somewhat later package development was the hockey-puck capsule, in which the diffusion is sandwiched between two heavy copper blocks, and held in place by a clamp which exerts a large compression force on the puck. This arrangement is hermetically sealed, compact and provides two paths for heat transfer and is still in use

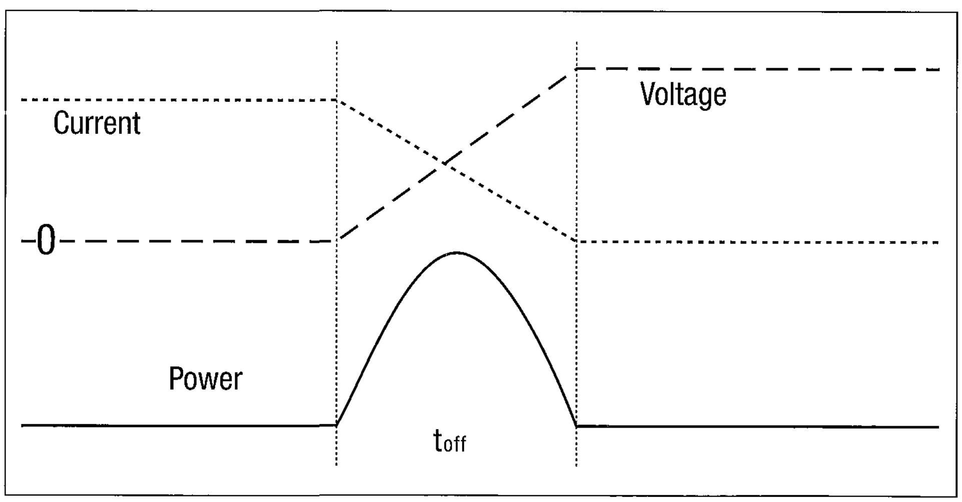

Figure 7. Simple illustration of switching power loss in a semiconductor

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today. The biggest disadvantage of the puck package is that the heat sink is electrically connected to the device, which requires that the heat sink be insulated from ground and other parts of the circuit.

In the early 1980s isolated base devices became available. In this package, one or more diffusions (active part of the device) are soldered or clamped down to a thin insulating layer of aluminum oxide or aluminum nitride. Below the insulating layer is a thick copper plate which is mounted on the heat sink. The advantage of this arrangement is that only one grounded heat sink is necessary, and all the devices are isolated from one another. Even though there is the additional thermal impedance, the isolated base module permits so much more design flexibility than the puck-type package that is has become extremely popular. All low-voltage drives today are constructed with isolated base modules. Still, GTOs, IGCTs and IEGTs and even some high-power IGBTs are made in the puck package. (Some products are available in both packages.) The module is not quite as well sealed as the puck and because the top connections are made with bond wires, the module can fail in an open-circuit state. In series strings of devices it is important that the device fail short to maintain a current path. At this writing, the isolated base module is more commonly used to construct medium voltage drives, although there are puck-based circuits available. Recent advances in the wire bonding techniques and baseplate material have improved reliability to the extent that module based inverters are now used in railway traction application. Figure 8 shows a puck device and an isolated base device.

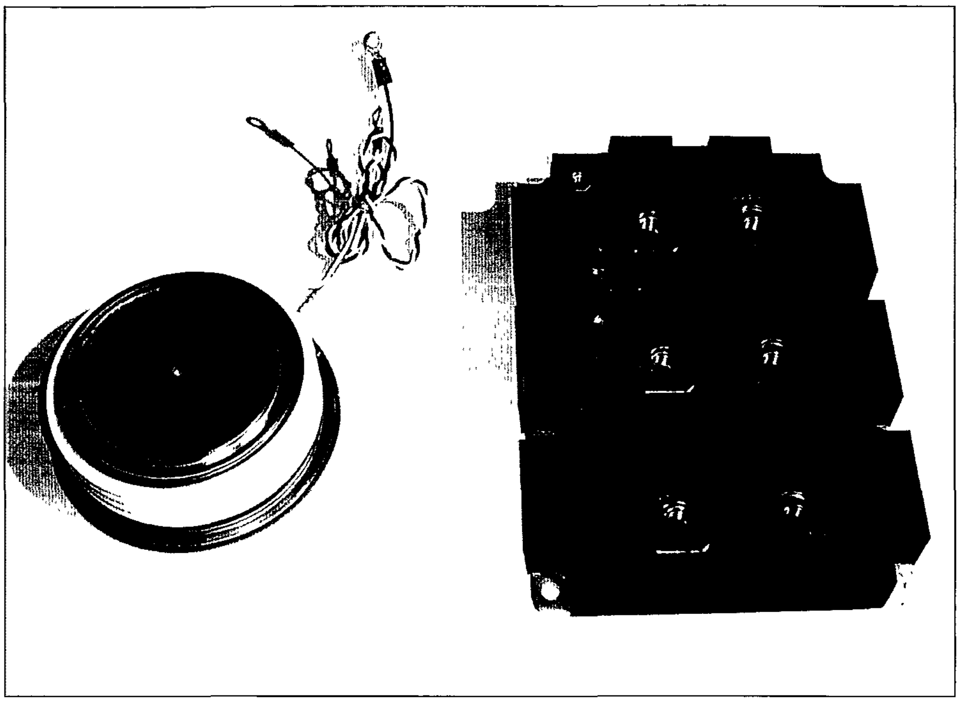

Figure 8. Picture of puck housing (left) and isolated base device

Device type	Maximum voltage	Maximum current	u sec off time	Peak gate power	"ON" voltage	$/kVA
Diode	7000	10000	50	N/A	1.0	Low
Thyristor	7000	10000	10-	2 W	1.25	Low
GTO	6000	4000	10 – 50	12kW	3.5	High
Transistor	1400	1800	3 – 5	20 W	2.5	Med
IGBT	3300	1200	1	3 W	3.5	Med
IGCT	6500	4000	2 – 3	45kW	2.0	Med

Figure 9. Summary of approximate device rating limits for commercially available units

The diode, thyristor, IGBT and IGCT form the device technology base on which the solid-state variable speed drive industry rests today. There are other device technologies and enhancements in various stages of development that may or may not become significant depending on their cost and availability in large current (> 50 A and high voltage 1,000 V) ratings. These include: two silicon carbide semiconductors, three variants of the four-layer switch such as the MTO (MOS turn-off thyristor) and MCT (MOS controlled thyristor). We should expect new switches to come along and significantly improve on the devices currently in use. While the type of semiconductor device is not necessarily the most important issue to a user, in general the newer devices provide better drive performance. See Figure 9 for a summary of device rating limits.

AC Variable Frequency Drives

The impact of new solid-state switching devices has been extremely significant on AC variable frequency drives and will likely continue. Solid-state variable speed drives have been developed and marketed for wound-rotor induction motors (WRIMs), cage-type induction motors, and synchronous motors.

Historically, WRIM-based variable speed drives were in common use long before solid-state electronics. These drives operate on the principle of deliberately creating high-slip conditions in the machine and then disposing of the large rotor power that results. This is done by varying the effective resistance seen by the rotor windings, and thence the name of slip-energy-recovery drives. But, the WRIM is the most expensive AC machine. This has made WRIM-based variable speed drives noncompetitive as compared with cage induction motor (IM) drives or load commutated inverters using synchronous machines. Except in developing countries, the WRIM has become a casualty of the tremendous progress in AC variable frequency drives as applied to cage induction motors and will not be discussed further.

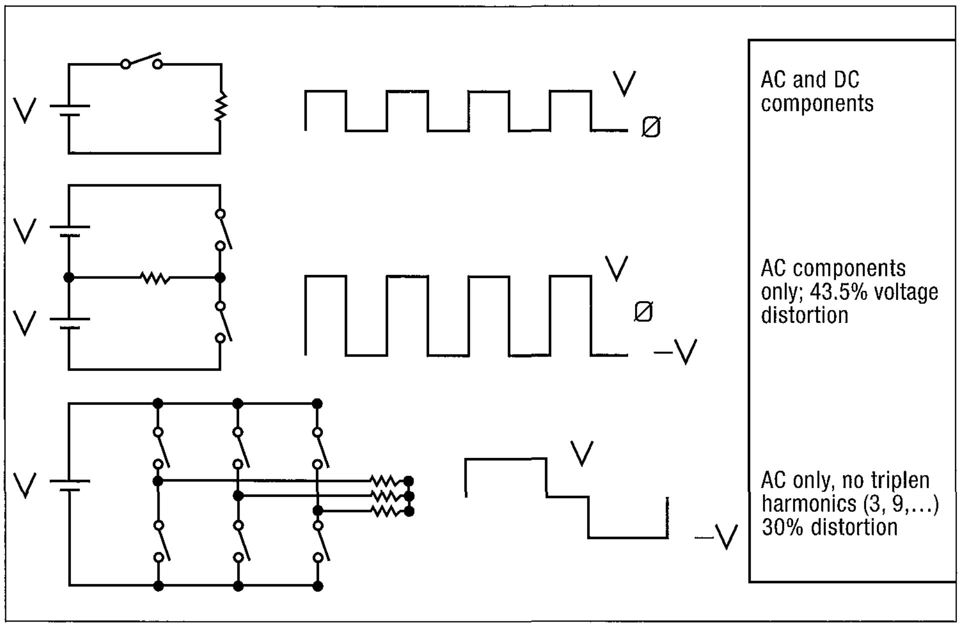

Figure 10. Waveforms of PWM voltage-fed VFD
Top—Motor current/Bottom—VFD voltage output, line-to-line

PULSE-WIDTH MODULATION

A basic concept in VFD is the method of creating the output waveform. Since the switching devices must be either on or off, the option of analog replication of a sine wave like a hi-fi amplifier is not open. Even if we used devices as linear amplifiers, the efficiency would be unacceptably poor. Many years ago it was discovered that a useful AC output could be obtained by using a variable amplitude DC link voltage and a very simple switching scheme.

Figure 10 demonstrates that by using the bridge circuit with 3 poles and switching the devices in a pole in a 50% duty cycle complementary manner (*i.e.,* one is on while the other is off), and phase shifting the modulation 120 degrees between poles, we can get an AC waveform with only 30% distortion. This was extended to using a fixed voltage and more elaborate pulse patterns to control both the amplitude and the frequency of the output. The technique of controlling the output amplitude and harmonic content by clever pulse patterns is known as pulse-width modulation. This ubiquitous method is used in both voltage-fed and current-fed circuits. The more switching events or pulses introduced into a cycle, the better the waveform becomes.

The process effectively eliminates low-order harmonics in the output by moving them to a higher frequency where they can be more easily filtered. In the case of voltage-fed circuits, the filtering is provided by the leakage inductance of the motor. The limiting factor is how much device energy is dissipated per switching event.

INDUCTION MOTOR VARIABLE SPEED DRIVES

Because the squirrel cage induction motor is the least expensive, least complex and most rugged electric machine, great effort has gone into drive development to exploit the machine's superior qualities. Owing to its simplicity, it is the least amenable to variable speed operation. Since it has only one electrical input port, the drive must control flux and torque simultaneously through this single input, as there is no access to the rotor circuits. In an induction motor of the power crossing the air gap, the slip portion is dissipated as heat in the rotor, and one-slip comes out the shaft as mechanical power. The rotor power dissipation raises its temperature, so very low-slip operation is essential. Induction motor variable speed drives in the past have had the greatest diversity of power circuits. These circuits can be divided into two broad categories: *current-fed and voltage-fed.*

CURRENT-FED VERSUS VOLTAGE-FED CIRCUITS: TWO BASIC TOPOLOGIES

Voltage-fed and current-fed refer to the two basic VFD strategies of applying power to the motor. In Europe, these are called voltage-impressed and current-impressed, which is a much clearer description. In voltage-fed circuits, the output of the inverter is a voltage, almost always the DC link voltage or its inverse. The motor and its load—not the inverter—determine the current that flows. Usually, these drives have diode rectifiers on the input. The main DC link filter is a capacitor. In current-fed circuits, the output of the inverter is a current, usually the DC link current or its inverse.

The motor and its load—not the inverter—determine the voltage. Usually these VFDs have a thyristor converter input stage and the DC link element is an inductor.

Today, voltage-fed VFDs use a rectifier bridge or multipulse bridges, or occasionally active front ends. This gives them consistently high P.F. and minimum high-order harmonics. The reactive power needs of the motor come from the capacitor and are not reflected to the line. But, the DC link electrolytic capacitors can be a reliability and lifetime issue. Energy stored in the link is very high compared to the CSIs, and a fault in the inverter can lead to very high currents. The motor's inherent inductance can be conveniently used to filter a PWM voltage wave. On the other hand, very fast wavefronts have become a concern to motor designers and users. In a PWM voltage-fed circuit, the output switches are controlled such that both the amplitude and frequency of the output are regulated. The most common approach in current-fed inverters is to use a thyristor converter on the line side to control the current and thus the amplitude of the output. The output switches control only the frequency of the output. The input power factor is the load power factor times the PU speed. The reactive power demand of the motor is passed back to the line. High order harmonics are present due to the high di/dt. Link energy storage is relatively low and the DC link reactor provides immunity to faults and grounds. Since the current is regulated, inverter faults do not cause high currents. The motor current cannot change instantaneously, so all the CSI circuits require a capacitive filter on the motor to absorb the high di/dt of the inverter.

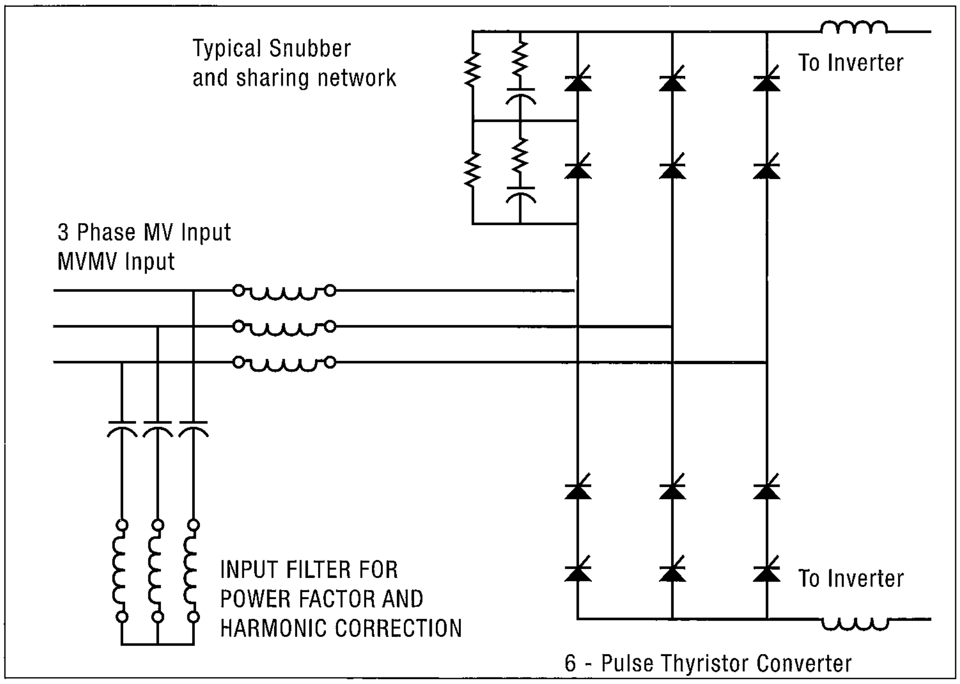

Figure 11. Six-pulse thyristor converter with series devices

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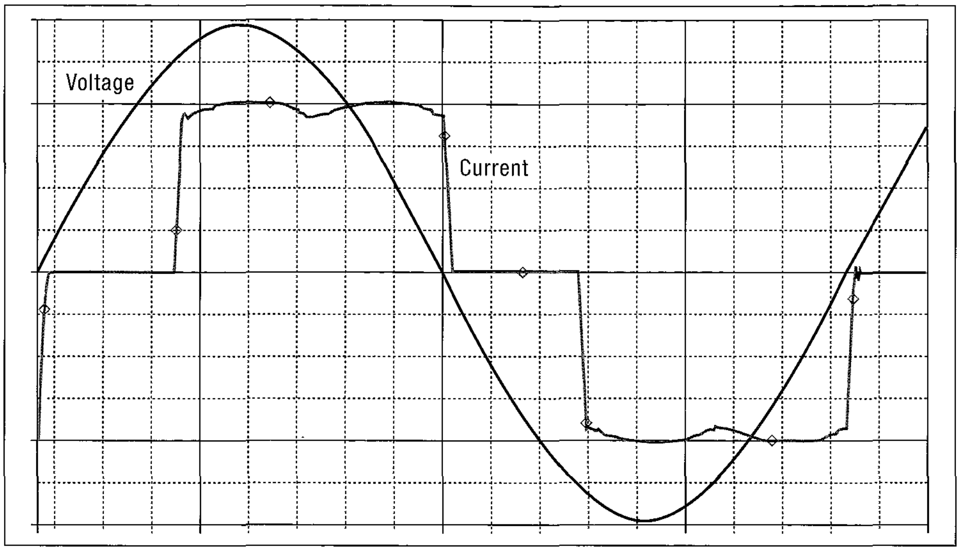

Figure 12. Typical input current to a 6-pulse thyristor converter

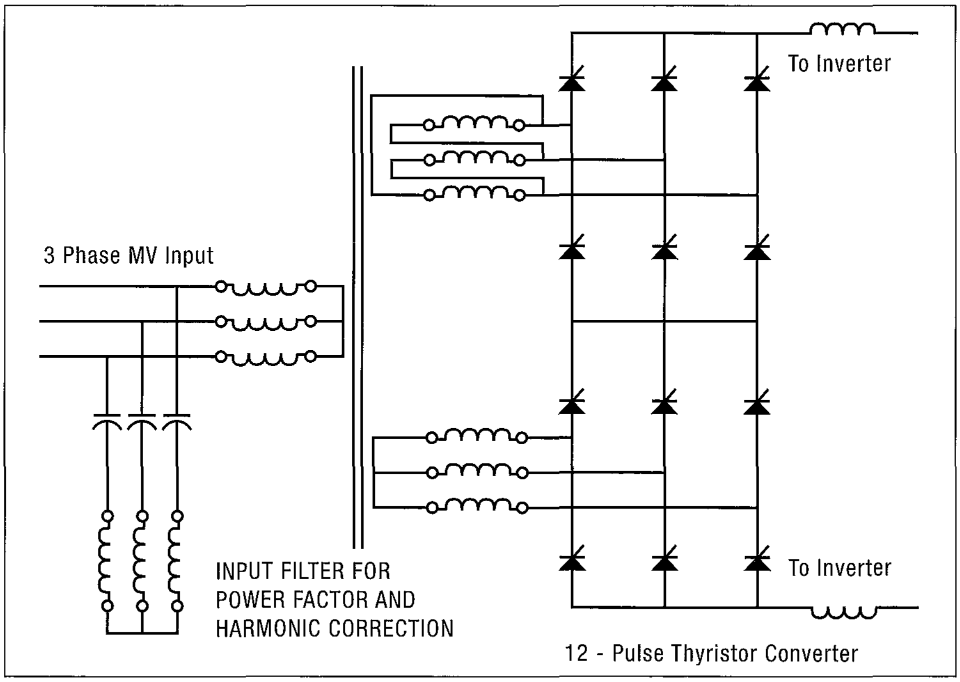

Figure 13. Twelve-pulse thyristor converter with phase shifting input transformer

LINE-SIDE CONVERSION

In a variable frequency drive, the nature of the line-side converter circuit between the utility and the DC link determines the input properties such as utility harmonics and power factor. This is done in two ways.

Beginning with input conversion of the current-fed circuits, Figures 11 and 13 illustrate two common alternatives for the line-side converter of both of the current-fed circuits. In Figure 11, the converter is a single three-phase bridge fed directly from the 4kV line with a line reactor. This arrangement requires two 5kV symmetrical thyristors in series to withstand the peak voltage of 5600 volts, plus some derating for imperfect sharing of voltage (In the usual industry practice, devices are applied at 50-60% of their voltage rating). This is perhaps the simplest and cheapest circuit, but it has the poorest harmonic performance. The input current spectrum has 20% fifth and 12% seventh harmonics and the rapid commutation rate results in significant harmonic components out to the 35th and beyond as shown in Figure 15.

This configuration, unless equipped with a filter, could not be recommended for any application above 1 MW in which the user has concerns about input power quality.

Applying thyristors, or any other semiconductor device in series, necessitates some means of assuring that the devices evenly share the voltages during switching and during blocking. As the leakage currents are generally unequal (one must assume some will be maximum specified and others zero leakage), some parallel resistance low enough to cause a parallel current which swamps out the device leakage is used. These sharing resistors can be minimized by matching devices, but always represent an additional circuit complication and extra power losses. Assuring sharing during turn-on and turn-off are more difficult. The gate drive circuits have to have short and closely matched propagation delays. During turn-off the difference in recovered charge must be absorbed by the device snubber. Either a large snubber or matched devices must be employed.

Figure 13 shows two thyristor converters in series, fed from a transformer with two secondary (wye and delta) windings. Each converter must be able to produce about 2800VDC maximum, so the secondary voltages are about 2000 VAC. Conversion at this voltage is readily possible with one phase-control thyristor rated at 5kV. The advantages of this arrangement are that it eliminates series devices, raises the pulse number from 6 to 12 and it permits the transformer to support the converter common-mode voltage rather than apply it to the motor. (See Figure 15 for a bar chart of the characteristic harmonics of these

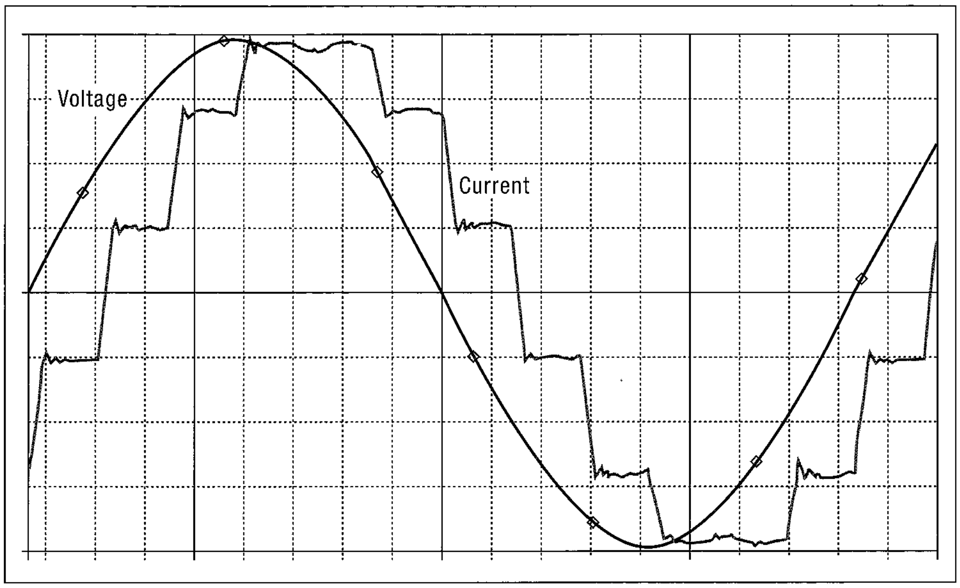

Figure 14. Input voltage and current waveforms for a 12-pulse thyristor converter

circuits.) Of course, the transformer has to be designed for the harmonic currents present in the primary and secondary currents. The main result of raising the pulse number from 6 to 12 is a dramatic reduction of the troublesome fifth and seventh harmonic currents. For thyristor converters the proportion of harmonic currents is very nearly 1/h of the fundamental, where h is the harmonic number. H = n*p +/- 1, where n is an integer and p is the pulse number. A very important point is that although only half of the harmonic spectrum is present in the 12-pulse case compared to the 6-pulse, those components of the 12-pulse spectrum are about equal to the 6-pulse values. Because of the rapid commutation rates, the high order harmonics are quite significant. This will be shown to be a significant difference compared to diode rectifiers. For the current-fed cases, the total current harmonic distortion at the converter input is approximately 30% for 6-pulse and 15% for 12-pulse. Neither figure approaches the strictest limit of IEEE-519 for the situation where the utility line short-circuit current is less than 20 times the total load current. Therefore, both circuits may require filtering to meet the 519 requirements, unless there is also a linear load of 3 to 6 times the drive rating.

The substantial quantity of higher-order harmonics can result in these SCR converters being the source of telephone interference. In that case, the filtering becomes dramatically more difficult.

In the current-fed circuits, the DC link voltage is equal to the motor rated peak line-to-line voltage times the actual load power factor at the operating point, times the PU speed. Therefore, in a centrifugal load, the DC link voltage drops rapidly as the speed drops, and the input converter must phase back to accommodate this effect. The phase-back results in a direct reduction in displacement power factor. Another aspect of the power factor issue is that since the load current flows directly through the link inductor into the converter (*i.e.*, the inverter current and the converter current must be identical), the reactive current requirement of the load is "passed back" to the line. There is no difference in the displacement power factor between the circuits of Figure 11 and Figure 13. See Figure 16 for the uncorrected P.F. of these converters versus stator frequency assuming a centrifugal load. In order to deal with the input power quality issues, these circuits are usually equipped with a substantial filter (~.3 PU) which corrects the power factor and absorbs some of the harmonic currents produced by the converter. This is shown in Figures 11 and 13 as a tuned branch. Fortunately, a fixed amount of reactive current compensation provides reasonably good P.F. over the usual operating speed range of a pump or fan (50%

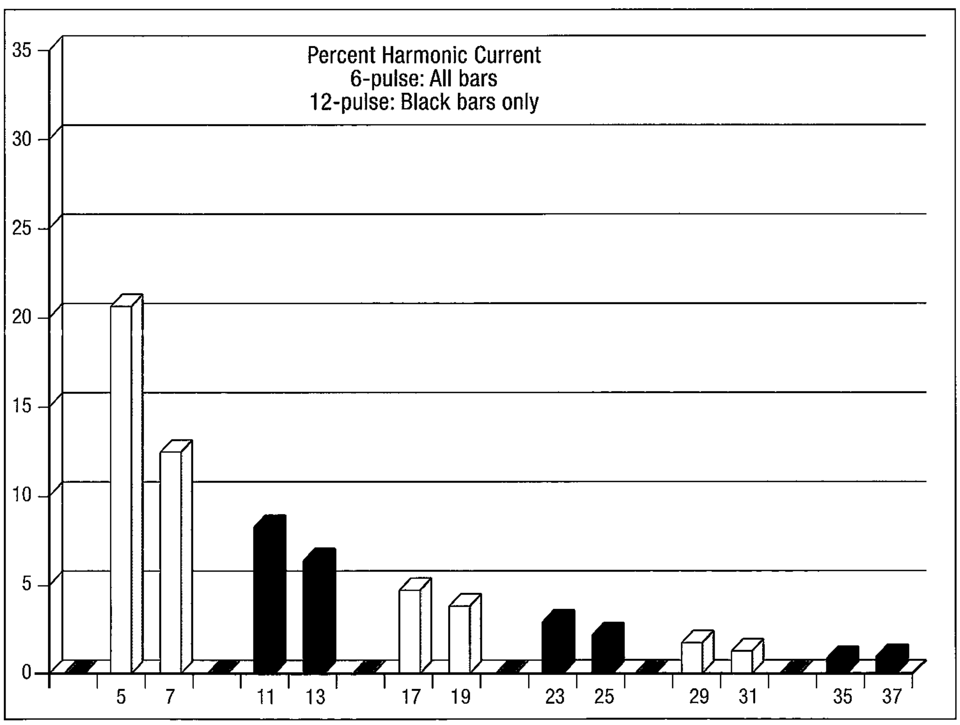

Figure 15. Harmonic current spectrum of a thyristor converter in 6- and 12-pulse circuits

to 100%)(Ref.1). In the case of the transformer with two secondary windings, a filter can be applied to both secondaries, thereby reducing the harmonic burden on the transformer, as well as lowering the total transformer fundamental current. Filters are more commonly applied at the transformer input.

In all the current-fed circuits, amplitude control of the output is achieved by controlling the DC link current with a regulator which manipulates the converter output voltage through phase angle control. Since the converter output voltage must track the inverter bus voltage to maintain the current, the phaseback angle is in a constant state of modulation. If the link inductor is small, this effect if aggravated, and then the conduction interval of the converter thyristors becomes unequal from cycle to cycle. This phenomenon results in non-characteristic input harmonics.

Although the DC link current always flows in the same direction, power flow from the motor to the line can be accommodated, because in that case the link voltage reverses polarity. Therefore, these input converters permit regeneration, and drives based on them are inherently four-quadrant. During regeneration, the input current waveform is the same as for motoring, but the angle of the current lies between 90° and 150° lagging the voltage.

Since the DC link is current controlled by a fast regulator in conjunction with the link inductor, fault protection downstream of the DC link is relatively easy.

An inverter commutation failure simply results in the converter having to phase back quickly to hold the current at its reference value. A very limited amount of fault current will flow. Having part of the link choke in each DC leg affords better protection. In that case, even a ground fault downstream of the link will result in only limited fault current.

A drawback of all power conversion circuits, but especially thyristor input circuits without isolation transformers, is that they will generate a large common-mode voltage, which appears on the motor winding-to-ground circuit insulation (Ref. 4,5). This phenomenon occurs because only two input lines are in conduction at a given time. Thus, the circuit downstream of the converter must assume a common-mode potential equal to the mid-point of the two conducting input phases. **This is not the neutral voltage, as would be the case if all three phases were uniformly loaded.** The common-mode voltage is a minimum at zero phaseback, but increases greatly as the phaseback angle increases, reaching a maximum at 90°.

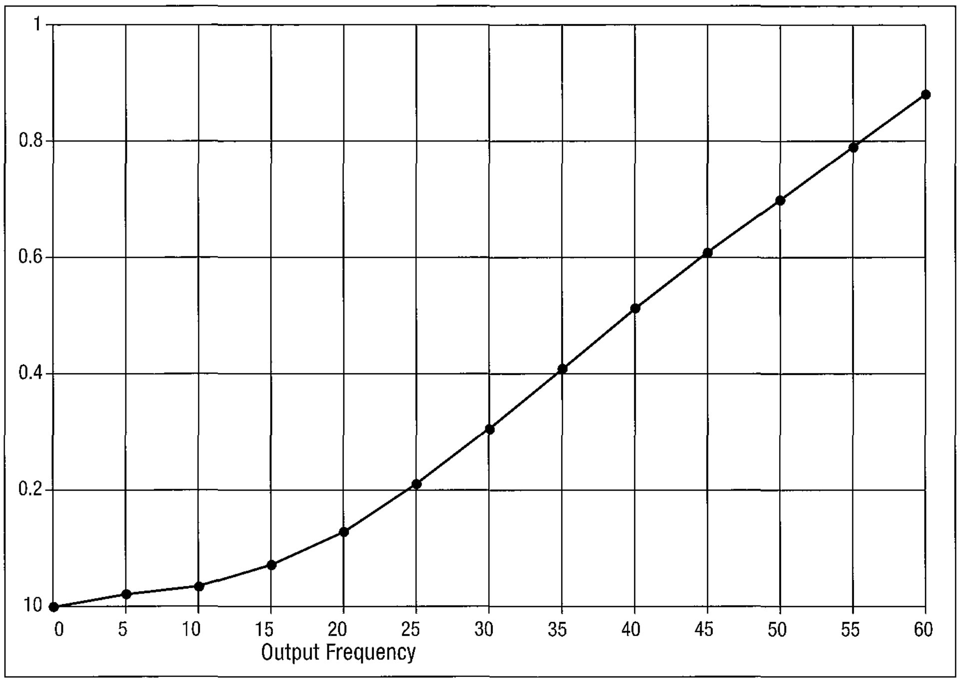

Figure 16. Uncorrected input power factor of current-fed circuits with centrifugal loads

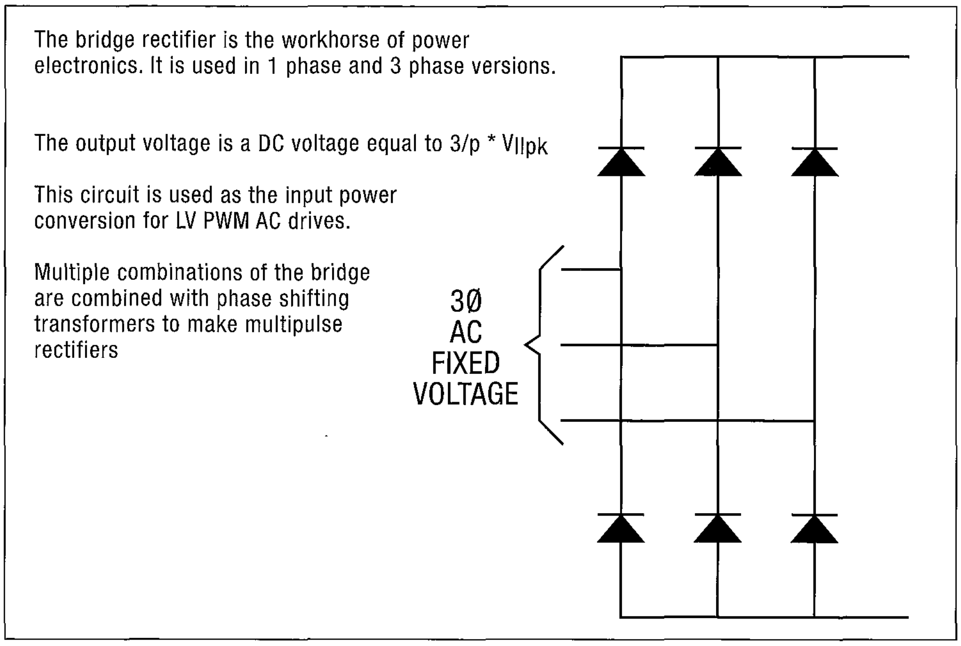

Figure 17. The diode rectifier bridge circuit

VOLTAGE-FED LINE CONVERSION

The behavior of rectifier circuits is somewhat different since they use uncontrolled devices-diodes which conduct as soon as the voltage becomes positive. The bridge rectifier is the most common circuit.

In this circuit, Figure 17, the positive bus is at the potential of the most positive line voltage, while the negative bus is at the potential of the most negative line voltage. (It's like an auction—the highest potential line wins.) Because of this action, the bridge is quite sensitive to unbalanced input voltage; the high line-to-line voltage tends to monopolize the current and third harmonic currents are drawn.

As seen in Figure 18, the input current is quite distorted, with large fifth and seventh current harmonics. But since the rate of change of current is low, the higher order harmonics are smaller than in the thyristor converter.

This circuit is used as the building block for multi-phase arrangements to reduce the current distortion. The input displacement power factor is uniformly high. But, this circuit cannot return energy to the line as can the controlled bridge.

AC and DC side inductors are frequently used to reduce the input harmonic current. The line-side inductor slows the commutation and widens the current pulses, reducing the 5th and 7th harmonics. The current waveform is very sensitive to the line-side inductance. This is arguably the most basic and inexpensive power conversion unit.

In Figure 19 we have two 6-pulse bridge rectifiers in series. Each bridge is fed from a separate 2400 VAC winding on the transformer. The DC link voltage is nominally 6800 VDC, with a midpoint established at the center of the capacitors. (Of course, one could use series rectifiers and operate directly from 4160 volts in a single 6-pulse circuit, if they didn't care about the harmonic consequences.) A DC link choke may be used between the rectifier and the capacitor, which will reduce the 5th harmonic current. However, this has the drawback of exposing the rectifiers to voltage transients on the input line. If the DC link inductor is not present, voltage transients are converted into current transients by the transformer reactance. These are much less likely to cause a rectifier failure than a voltage transient. A big advantage of this arrangement is that the 5th and 7th harmonic currents from the two bridges cancel in the transformer and are not present in the transformer primary current. Since uncontrolled rectifiers are used, the displacement power factor is nearly unity. Therefore, these input converter arrangements have inherently high power factor at all operating conditions and P.F. correction is unnecessary.

In voltage-fed circuits, the inverter current does not flow exclusively into the converter owing to the shunt path of the DC link capacitance. The reactive power requirements

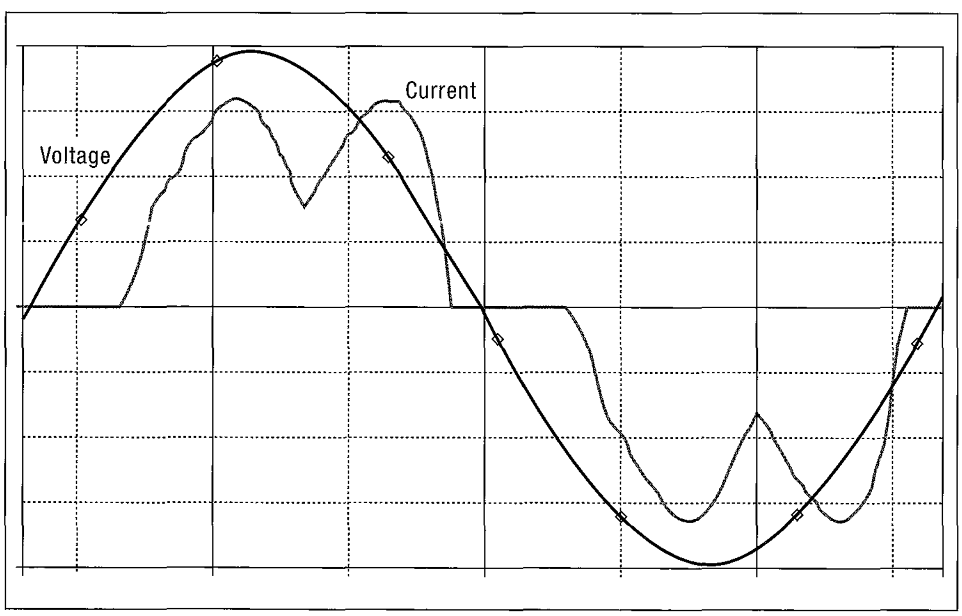

Figure 18. Typical input current waveform to a 6-pulse diode rectifier

of the load are supplied by the inverter, using the DC link capacitor as storage, since the average power of reactive currents is zero. So the input converter and transformer of a voltage-fed drive need only deal with real power requirements of the load, not the reactive component.

This results in higher efficiency at reduced speed. Conversely, since the DC link voltage is essentially constant, the output amplitude control must be achieved by the inverter via PWM strategies.

The input harmonic properties depend on the values of the AC side reactance and the DC link choke. As the rectifiers undergo commutation when the voltages a minimum, di/dt's are very low compared to thyristor converters.

A large AC side reactance further slows the commutation rate and minimizes higher order harmonics, while the DC link reactor is mostly effective at reducing 5th and 7th. See Figure 21 for a bar chart of the typical harmonic spectrum for the uncontrolled rectifier circuit assuming .05PU commutating reactance. In order to attain compliance with the most stringent category of IEEE-519-1992, 5% ITHD, it is necessary to add 0.3PU kVA of harmonic filter to the circuit of Figure 17. It is possible to dispense with the DC link choke by using other ways to deal with the augmented 5th and 7th currents which result.

Because of the rapid fall-off of the harmonic currents as compared to the thyristor converter (compare Figure 14 and Figure 20), a 12-pulse rectifier comes much closer (~7% ITHD) to meeting the most stringent current distortion IEEE-519-1992 limit of 5%.

Most neutral-point clamped circuits use this configuration, but occasionally there are 18-pulse circuits in difficult cases. An 18-pulse diode rectifier with suitable line-side reactance will attain 5% or less current distortion.

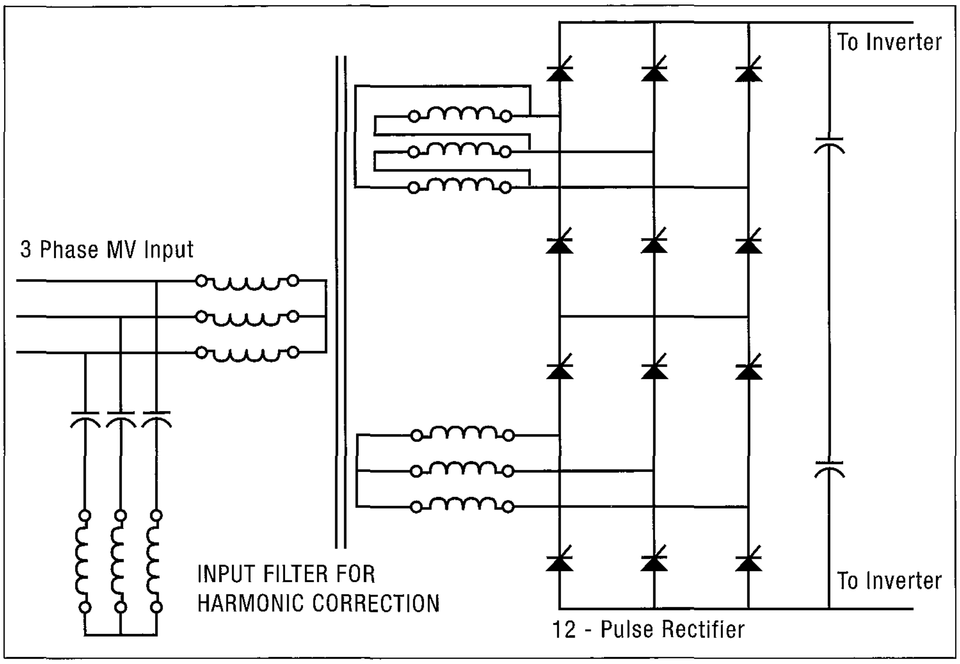

Figure 19. Twelve-pulse rectifier—bridges in series

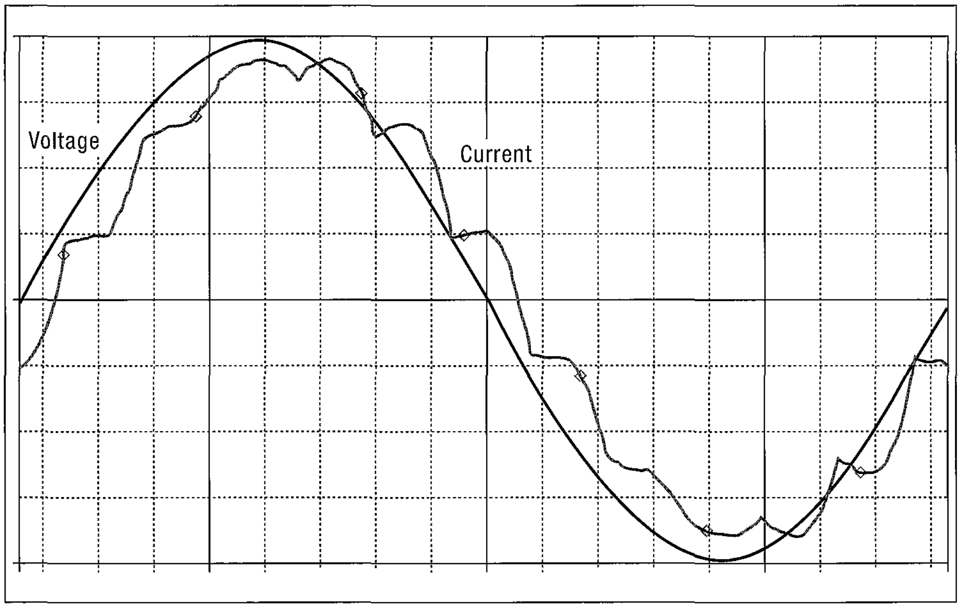

Figure 20. Typical input current waveform to 12-pulse rectifier

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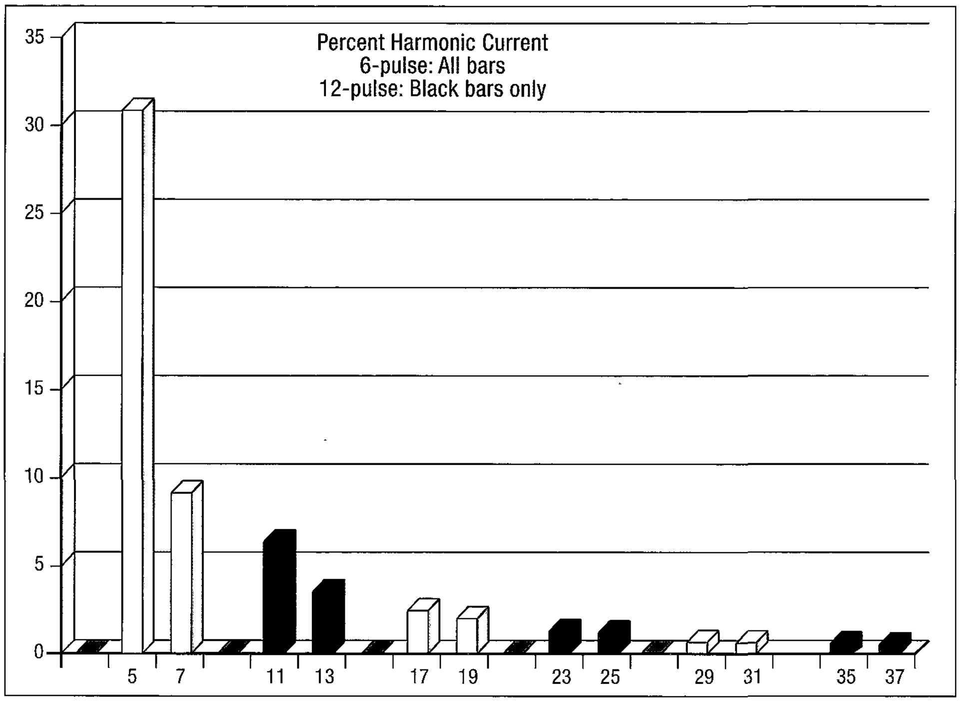

Figure 21. Input harmonic current spectrum of a rectifier bridge

ACTIVE FRONT END LINE-SIDE CONVERSION

Another way to convert the utility power to DC power for the link, frequently called the "active front end," has emerged as practical with the decrease in cost of devices. It uses six fully-controlled (that is, you can turn them off/on at will) switches (GTOs, IGCTs, or IGBTs). In this circuit, the switches are controlled with a PWM technique to generate a sinusoidally-modulated voltage at the AC input to the bridge. There is an inductance and filter connected between the AC input to the bridge and the utility.

By adjusting the amplitude and phase of the modulated voltage, the user can control the amount of current flowing into the bridge and also its phase. Therefore, this circuit can make power flow in either direction (permitting four-quadrant operation) and at any desired power factor. The PWM modulation process produces voltage harmonics at high frequencies.

The filter prevents these harmonic voltages from causing large harmonic currents into the utility. Of course, the big disadvantage is that the cost and complexity of the fully-controlled switches is much higher than that of diodes. Building such a circuit for current-fed topologies is also possible.

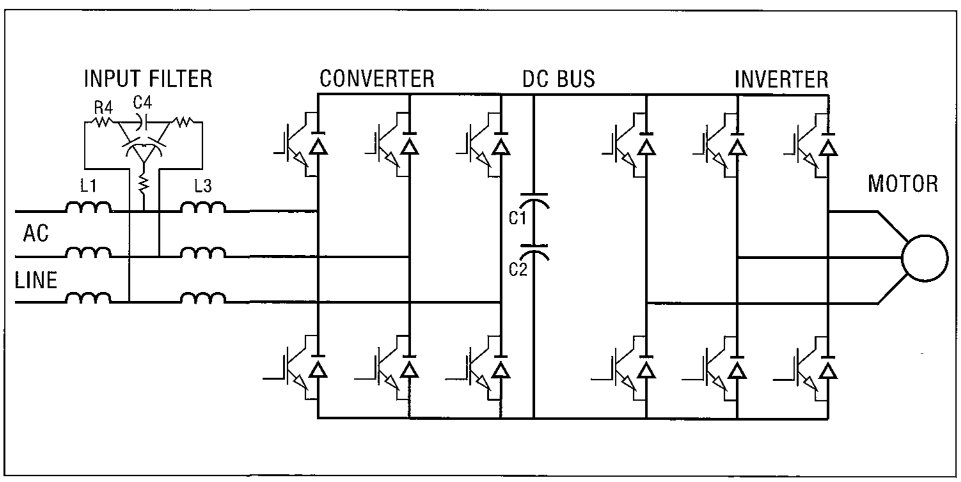

Figure 22. VFD with an active front end

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DRIVE CONTROL TECHNOLOGY

Parallel to the development of power switching devices, there have been very significant advances in hardware and software for controlling variable speed drives. These controls are a mixture of analog and digital signal processing.

The advent of integrated circuit operational amplifiers and integrated circuit logic families made possible dramatic reductions in the size and cost of the drive control, while permitting more sophisticated and complex control algorithms without a reliability penalty. These developments occurred between 1965 and 1975. Further consolidation of the control circuits occurred after 1975 as large-scale integrated circuits (LSI) became available. In fact, the pulse-width modulation (PWM) control technique was not practical until the appearance of LSI circuits because of the immense amount of combinatorial logic required. Clearly, the most significant advance in drive control has been the introduction of microprocessors into drive control circuits. The introduction of cheap and powerful microprocessors continues to expand the capability of drive controls. A modern drive should have most of these features. The performance enhancements include:

1. More elaborate and detailed diagnostics owing to the ability to store data relating to drive internal variables, such as current, speed, firing angle, etc.; the ability to signal to the user if a component has failed.
2. The ability to communicate both ways over industry standard protocols with the user's central computers about drive status.
3. The ability to make drive tuning adjustments via a keypad or over an Ethernet link with parameters such as loop gains, ramp rates, and current limits stored in memory rather than potentiometer settings.
4. Self-tuning and self-commissioning drive controls.
5. More adept techniques to overcome power circuit nonlinearities.

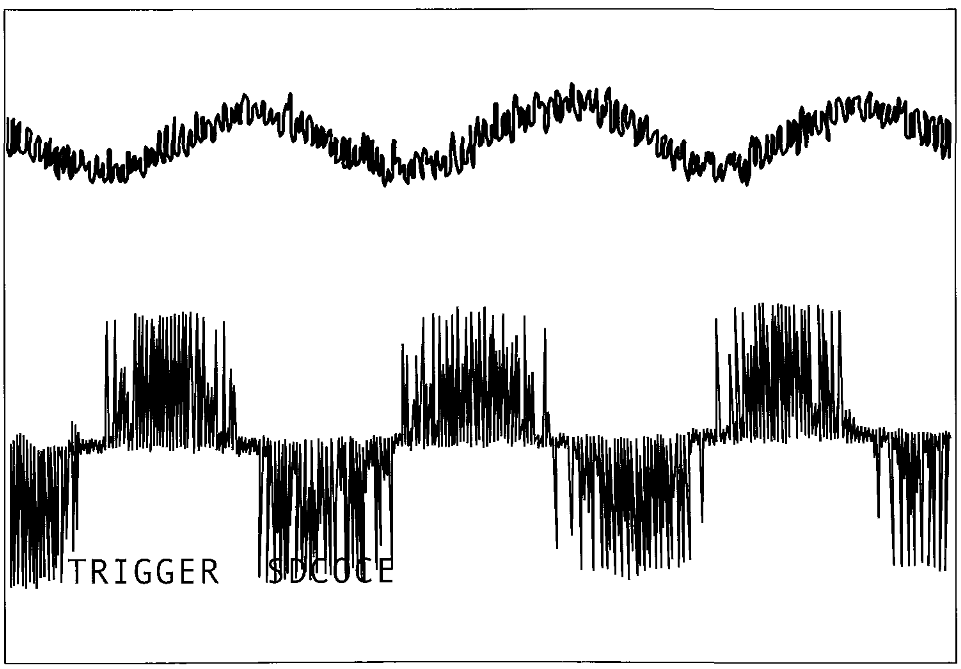

Figure 23. Generation of an AC output by switching a DC voltage

Medium Voltage Variable Frequency Drives

Understanding the details of the machine-side converters (inverters), requires an understanding of the needs of the "load"—the motor. AC machines have mostly been designed for operation on the utility, where the voltage is a reasonably good sine wave. So, the efforts of VFD manufacturers have been directed toward building a drive with an output close enough to the utility voltage to permit the motor to operate satisfactorily, preferably without derating and without a shortening of lifetime. The output of present-day medium voltage VFDs is not perfectly sinusoidal. First we need to avoid low-order voltage harmonics which will raise the motor rms current without creating additional torque, and which also cause torque pulsations. In a properly designed motor, with sinusoidal currents, the output torque is smooth with time. This is not the case when harmonic currents are present. Generally, the total harmonic distortion of the motor current should be less than 5%. Second, the motor groundwall insulation is not intended to cope with a continuous offset of the winding voltage from ground (common-mode voltage), so we need to keep the common-mode voltage to a low value. Third, the motor coil insulation was originally designed to deal with the very low dv/dt of the utility voltage, with infrequent transients superimposed. So, the drive output voltage needs to avoid subjecting the motor to repetitive transients with high dv/dt, (step functions) which may cause excessive turn-to-turn voltage. Fortunately, the coils in a medium voltage motor are much better designed for withstanding this stress than a low-voltage motor coil. This is because the medium voltage motor coils are form-wound with conductors insulated by varnish and by tape. The geometry of the coil is always the same and the start turn does not lie on the finish turn. By contrast, in a low voltage motor, the coils are usually random-wound so the start and finish might lie adjacent to one another. Also, there's usually no tape on the turn-to-turn insulation; it is only the varnish. The dv/dt limit for a medium voltage motor is 1000V/us.

For drives rated 2300VAC and above on the output, there are a number of choices of design of both current and voltage fed types.

1. The load-commutated inverter (LCI) (Figure 24)
2. The filter-commutated thyristor inverter (Figure 27)
3. The current-fed GTO/SGCT inverter (Figure 29)
4. The neutral-point-clamped inverter (Figure 31)
5. The multi-level series cell VFD (Figures 33, 34)
6. The cycloconverter (Figure 35)

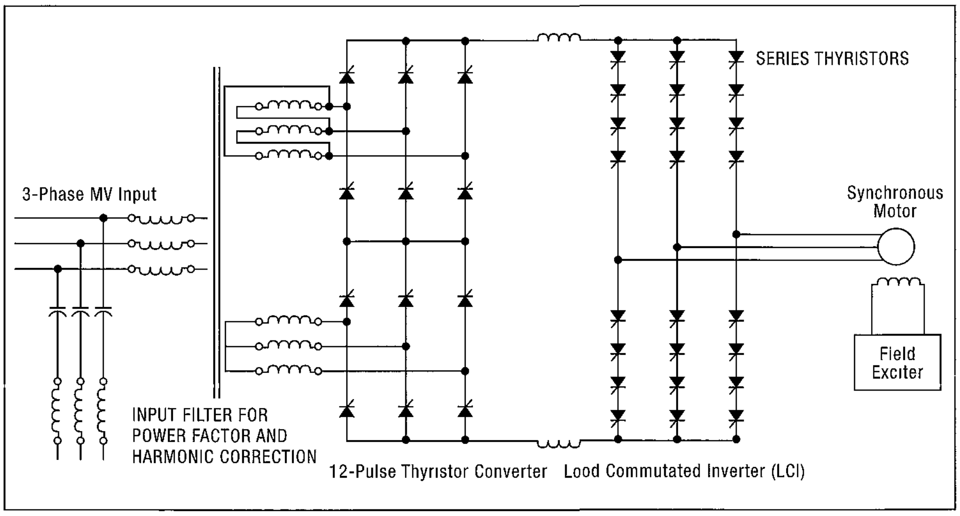

Figure 24. Load-commutated inverter (LCI)

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LOAD-COMMUTATED INVERTER

As shown in Figure 24, the load-commutated inverter is used with a synchronous machine. The LCI uses two thyristor bridges, one on the line side and one on the machine side. All the thyristors are "naturally" commutated. Commutation refers to the process of changing the current from one switching device to the next. In natural commutation, the utility line or some other low-impedance AC-voltage source provides the commutation energy. The synchronous motor acts like a utility line since its counter EMF commutates the machine-side converter. The machine-side converter operates exactly like the line-side converter, except the phase back angle is about 150°. The machine naturally applies reverse voltage to an off-going device before the next thyristor is gated. This imposes special design criteria on the synchronous motor. It must be able to operate at a leading power factor (0.9) over the speed range, it must have low enough leakage inductance to quickly commutate the thyristors, and it must be able to withstand harmonic currents in the damper windings. The requirement for the machine to always operate with a leading power factor requires more field excitation and a special exciter compared with that normally applied to a synchronous motor. This also results in a reduction in the motor torque for a given current. The machine-side devices are fired in exact synchronism with the rotation of the machine, so as to maintain constant torque angle and constant commutation margin. This is done either by rotor position feedback or by phase-control circuits driven by the machine terminal voltage. Only RC networks for voltage sharing are necessary. The output current is very similar in shape to the input current (a quasi-square wave), which implies a substantial harmonic component. The harmonic currents cause extra losses in the damper bars, and they give rise to very significant torque pulsations. The drive is not self-starting due to the low machine voltage at low speeds. Therefore, the drive is started by interrupting the DC link current with the line-side converter in order to commutate the inverter thyristors. The line-side converter is regulated to control torque. A choke is used between converters to smooth the link current, since there is considerable voltage ripple on the line-side and machine-side buses. Due to the harmonic and torque pulsation issues with the 6-pulse arrangement, LCIs with 12-pulse inputs feeding two sets of windings on the motor became popular, as shown in Figure 26.

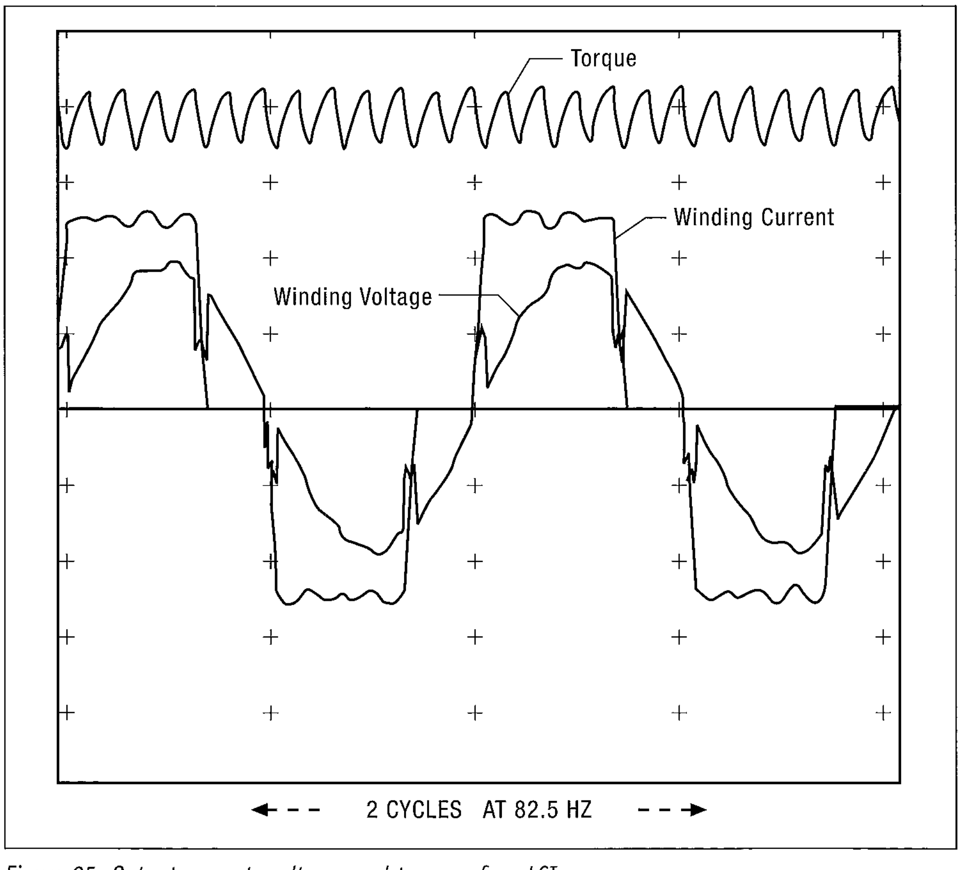

Figure 25. Output current, voltage, and torque of an LCI

Load commutated inverters (LCIs) came into commercial use in the late '70s and are used mainly on large medium voltage drives (1MW – 100MW). At these power levels, multiple series devices are employed (typically 4 at 4 kV input), and conversion takes place directly at 2.4 or 4 kV or higher. The efficiency is excellent, and reliability has been very good. Although they are capable of regeneration, LCls are rarely used in four -

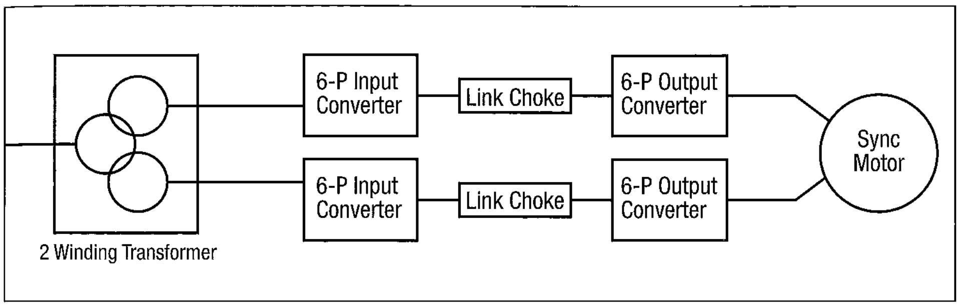

Figure 26. Twelve-pulse load-commutated inverter

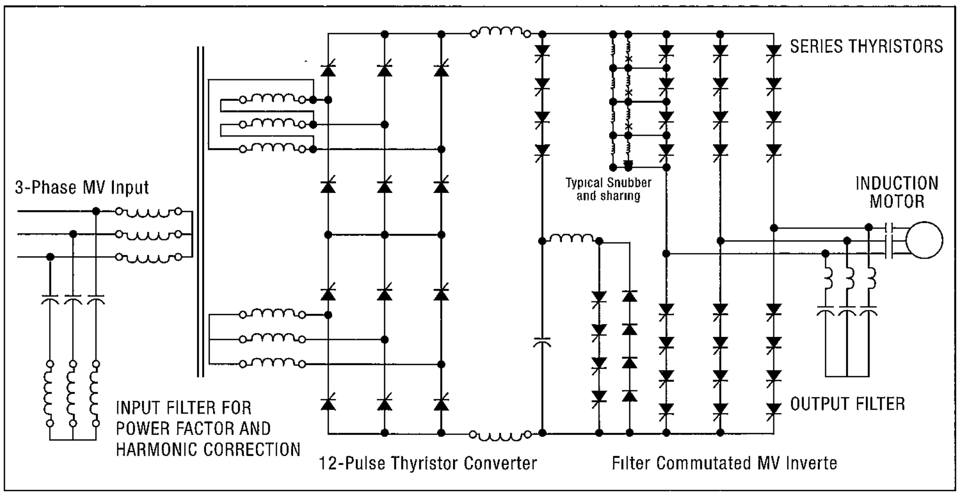

Figure 27. Filter-commutated medium voltage induction motor VFD

quadrant applications because of the difficulty in commutating at very low speeds where the machine voltage is very small. Operation above line frequency is straightforward. Despite the need for special synchronous motors, the LCI drive has been very successful, particularly in very large sizes, where only thyristors can provide the current and voltage ratings necessary. Also, many high-speed LCIs have been built. Now that self-commutated VFDs are available, the LCI is becoming less popular.

FILTER-COMMUTATED THYRISTOR DRIVE

In the circuit shown in Figure 27, the synchronous motor is replaced with an induction motor and output filter capacitor. The output filter capacitor is chosen to supply all the magnetizing current of the motor at about 50% speed. Above that point the load (machine and filter combined) power factor remains leading and the inverter thyristors are naturally commutated, that is, the voltage across the device is naturally reversed before the reapplication of forward voltage. In this mode of operation, the thyristor waveforms are similar to those in an LCI. The filter must supply (at a minimum) all of the reactive current requirements of the motor at full load, and is typically 1 PU of the drive kVA rating. In addition to the large AC capacitors, the filter requires some series inductive reactance to limit the di/dt applied to the inverter thyristors. Since the filter is capable of self-exciting the motor, a contactor is required to isolate the filter from the motor when the drive is off line. The large filter has the advantage of providing a path for the harmonic currents in the inverter output (which is a 6-step current), so that the motor current waveform is good near rated frequency. As the output frequency decreases, the filter becomes less effective and the motor current waveform deteriorates. The fundamental current into the filter increases with the square of the frequency up to rated voltage, since the voltage is also increasing with the frequency. Since the filter cannot provide commutation down to zero frequency, it is necessary to provide an auxiliary commutation means to get the drive started. This circuit acts on the DC link current, and is commonly called the diverter. When it is time to switch inverter thyristors, the DC link current is temporarily interrupted

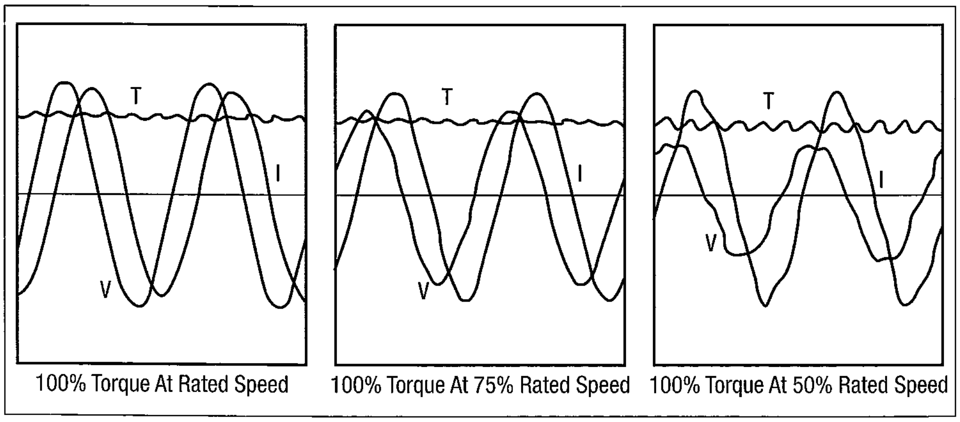

Figure 28. Output voltage, current and torque of a filter-commutated VFD

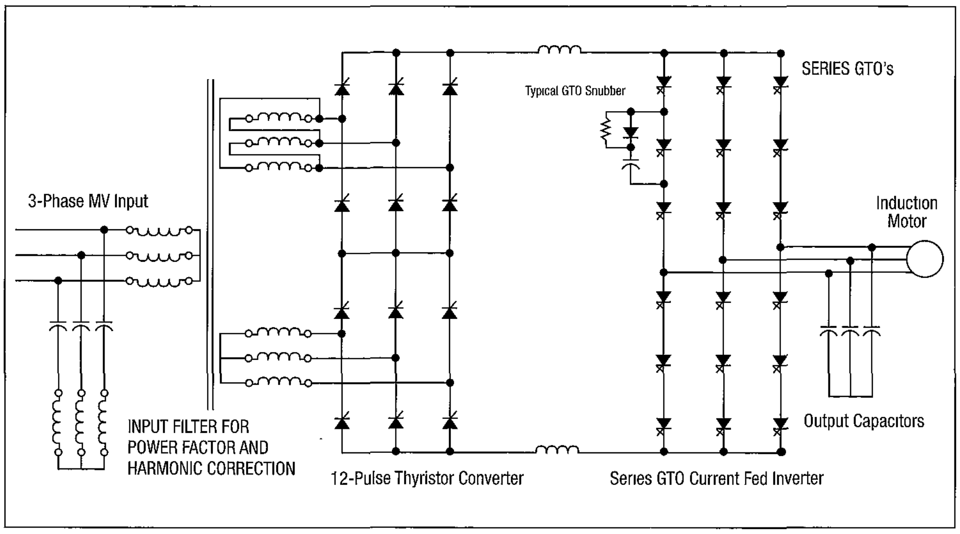

Figure 29. Current-fed VFD using GTO/IGCT inverter stage

(diverted), allowing the devices to recover. Then the next thyristor pair is gated, and DC link current is restored. The auxiliary circuit needs to be able to withstand full link voltage, and interrupt the rated DC link current for several hundred microseconds to permit the inverter thyristors to recover. (High voltage thyristors require long turn-off times as a consequence of design compromises in achieving high-blocking voltage.) Thus, the auxiliary commutation circuit is quite significant in rating. It is not usually intended for continuous operation, but only to get the speed up to the point where the filter commutation commences. The drive controller has two modes of operation. This circuit has been implemented using four 3kV thyristors in series per leg of the output bridge. (It is possible to add additional thyristors for redundancy.) Each leg of the bridge experiences the peak motor line-line voltage of about 6000 volts in both polarities, so the devices must have symmetrical blocking voltage. As in the input converter, the issue of voltage sharing during steady-state and switching arises. Combinations of device matching and/or RC snubbers are needed. Gate circuits for thyristors are simple and typically deliver 3 – 5 watts power, although they are designed for somewhat more. This approach has been most successful in those applications where the drive operates more or less continuously and in the range of 60 – 100% of rated speed.

CURRENT-FED GTO INVERTER

Another medium voltage bridge inverter circuit is shown in Figure 29. Here the output devices are GTOs (three 4kV units per leg will be required) which can be turned off via the gate. This reduces the size of the filter as compared to the filter-commutated inverter to perhaps 0.8 PU, but it does not eliminate it. Since the motor appears to be a voltage source behind the leakage reactance, it is not possible to commutate the current between motor phases without a voltage to change the current in the leakage inductance. When a GTO turns off, there must still be a path for the current trapped in the motor leakage inductance, which is provided by the capacitor bank. The capacitors resonate with the motor leakage inductance during the transfer of current. The choice of capacitor is determined by the permissible maximum voltage during commutation and the location of the resonance with the motor inductance. All current-fed VFDs need a "buffering" capacitor between the impressed current of the inverter and the inductance of the motor. Furthermore, if the capacitor bank exceeds .2 PU, the possibility of self excitation of the motor exists, unless precautions are taken, such as a contactor between machine and drive. Voltage-fed circuits do not require output capacitors, because the voltage across the leakage inductance can be arbitrarily changed. Since the capacitor bank is smaller than in the filter-commutated VFD, it does not provide as much filtering of the output

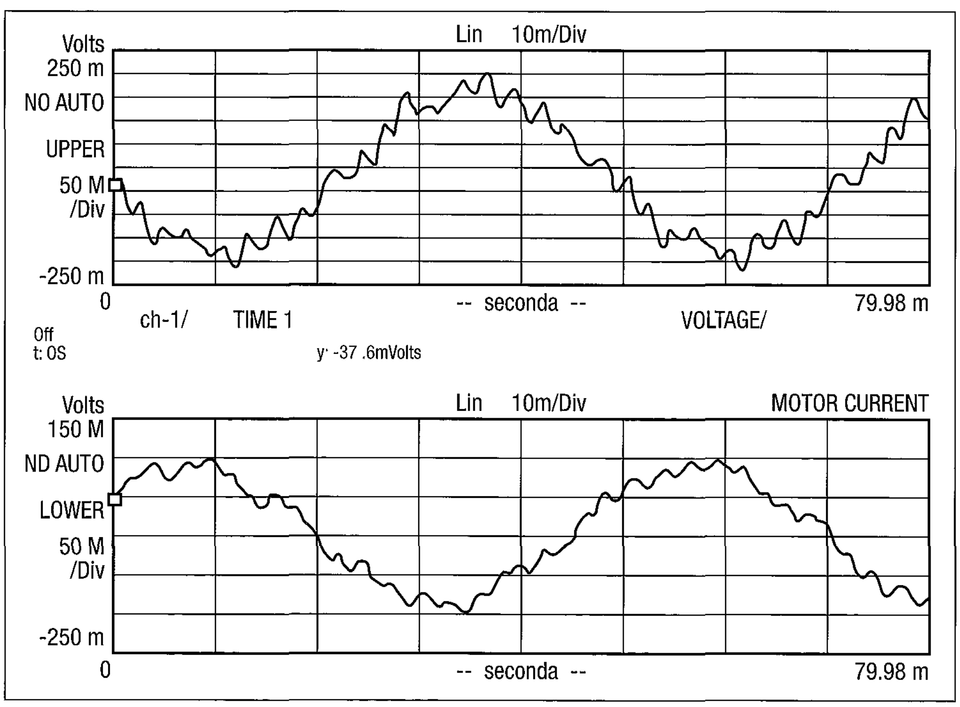

Figure 30. Output voltage (top) and current of a GTO/SGCT current-fed VFD

current. Motor current improvements are made by harmonic elimination switching patterns for the GTOs. At low frequencies, many pulses per cycle are possible and harmonic elimination is quite effective. But the GTO frequency limit of a few hundred Hz restricts harmonic elimination at rated frequency to the 5th and maybe the 7th.

This frequency limit is due to the nature of the GTO turn-off (and to a lesser extent, turn-on) mechanism. The device is turned off by extracting charge from the gate over a period of a few tens of microseconds and interrupting the regenerative turn-on mechanism. Near the end of the charge extraction period, the voltage across the GTO rises and the current begins to fall. During this time the device experiences extremely high internal power dissipation, which must be mitigated by the use of a large (1-5uF compared to .1uF for thyristors) polarized snubber located very close to the GTO. In that snubber, the capacitor is connected through a diode (the diode needs the same voltage rating as the GTO) to the GTO, so turn-off current can divert into the snubber, but the capacitor cannot discharge into the snubber at turn-on. The energy transferred to the snubber capacitor must be disposed of in some way so that the capacitor is discharged before the next turn-off. So GTOs typically have a minimum "on" time (10uS) and a minimum "off" time (100uS) to permit the internal switching heat to flow away from the junction and for the snubber to recover. Violation of the minimum time limits, or an unsuccessful turn-off attempt can result in destruction of the GTO. This limits the maximum switching rate with tolerable losses to a few hundred Hz. The GTO gate driver, in addition to providing a turn-on pulse comparable to the thyristor driver, must deliver a peak negative current of 1/5 to 1/3 the anode current in order to turn off the device. Thus, the GTO driver has a peak VA rating of 2 to 3 orders of magnitude higher than that for a thyristor, and perhaps ten times the average power requirement. This is an important factor in that all the gate power must be delivered to a circuit floating at medium voltage potential. The snubber losses can have a noticeable effect on part-load efficiency for a GTO drive. Some circuit implementations use patented energy recovery techniques to avoid efficiency deterioration, but these add serious complexity.

The snubber loss is proportional to the frequency and to the snubber capacitance, but to the square of the voltage. Those circuits need to use devices with a comparable voltage rating to the GTO. The design compromises in the metallurgy of the GTO results in a significantly higher forward drop (2.5 to 4 volts) than the conventional thyristor.

The device design is further complicated by the requirement for symmetrical voltage blocking in the current-fed topology. This circuit has benefited from the development of the IGCT and the SGCT. They perform much better than the GTOs in switching and low-forward drop and thus have improved the VFD performance considerably.

NEUTRAL-POINT-CLAMPED INVERTER

Despite the design issues of series GTO designs, they have also been used successfully in voltage-fed drives. Figure 31 illustrates such a circuit, the neutral-point-clamped inverter. There have been many of this type applied at 3300 volts output with 4.5kV GTOs, but the circuit has only recently been extended to 4kVAC, probably because of the improved properties of the IGCT. In the newer versions of this drive, the GTOs are replaced with IGCTs and IGBTs. One very important improvement is that the IGCT can operate with a very small snubber or no snubber at all.

In this 4kVAC output design, the total DC link voltage is 6kV, with a midpoint established at the center of the capacitor filter. Each leg of the bridge consists of two 6.5kV IGCTs in series. There are diodes in reverse across each GTO to permit motor current to flow back to the link, and still more diodes (same voltage rating as the GTOs/IGCTs) connecting the mid-points of the inverter legs back to the mid-point of the DC link. The total device count is 12 GTOs and 18 diodes (plus 12 more diodes in the GTO snubbers, if GTOs are used). The neutral-point-clamped inverter offers several advantages in those cases where series devices would be necessary anyway. First, the clamping diodes permit another voltage level, the DC link midpoint, at the output. This cuts the voltage step seen by the motor in half, and more important, creates another degree of freedom in eliminating output harmonics. Also, the clamping diode positively limits the voltage across any one device to half the link

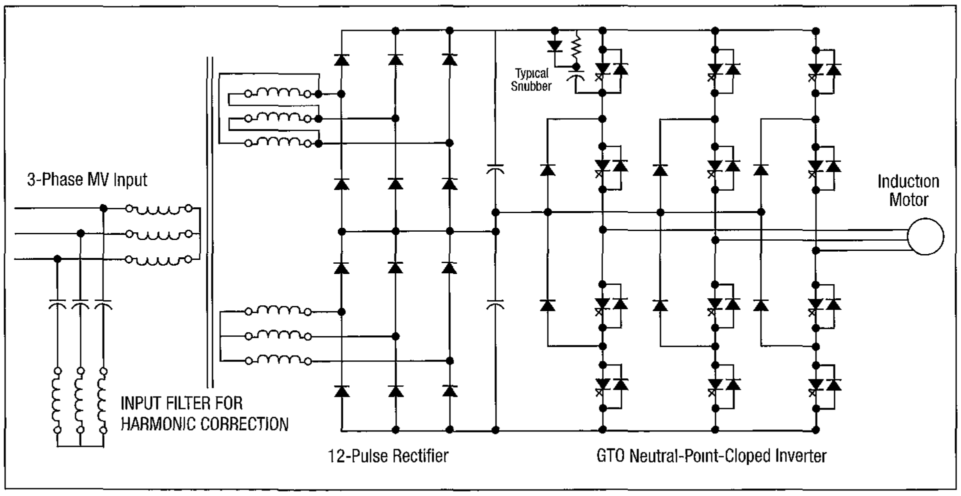

Figure 31. GTO/IGCT voltage-fed neutral-point-clamped inverter

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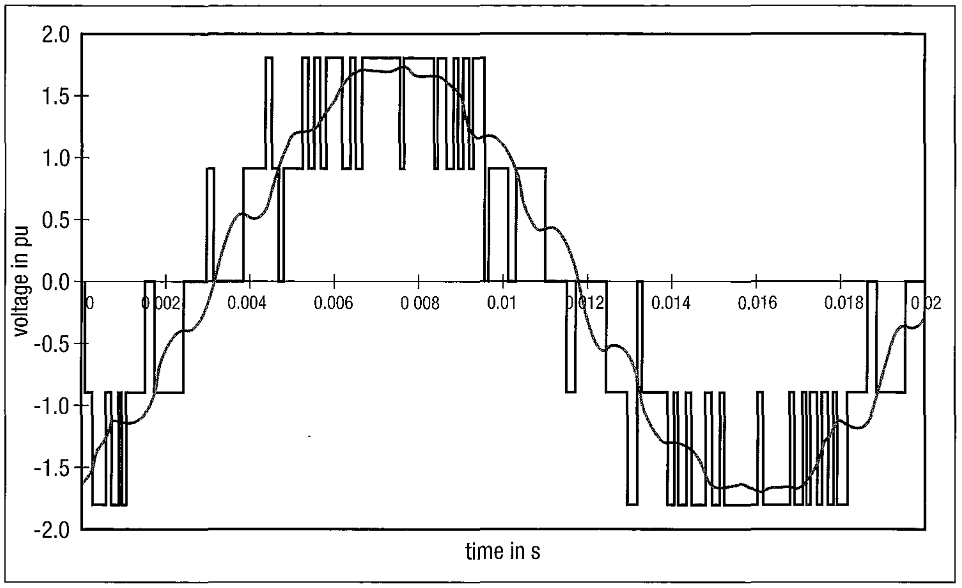

Figure 32. Raw and filtered output voltage waveform of an IGCT neutral-point-clamped inverter

voltage, enforcing voltage sharing without additional RC networks. Sometimes the NPC is equipped with an output filter to improve the motor waveform.

Since the switching devices in this circuit are never subjected to reverse voltage, using asymmetrical devices in which absence of reverse blocking is traded off for lower conduction and switching losses is preferable. Device protection during a short circuit can be a problem, as the GTO/IGCT can carry almost unlimited fault current like a thyristor. Unlike the current-fed circuits where fault current is limited, in the voltage-fed circuit, the DC link capacitor can source very large fault currents in the event of a short or a commutation failure. Protection schemes generally attempt to limit the rate of rise of fault current with an inductor and then turn off the devices before it grows beyond the safe turn-off level.

It is possible to use the NPC topology with IGBTs as the switching devices. As IGBTs are currently limited to 3300 volts, the IGBT NPC cannot yet reach 4kVAC output, but IGBT manufacturers are working on a 6kV IGBT. The concept of NPC can be extended to M-level inverters, although the number of diodes grows rapidly. Since each device is topologically unique, adding redundant devices would require twice as many, instead of just one more.

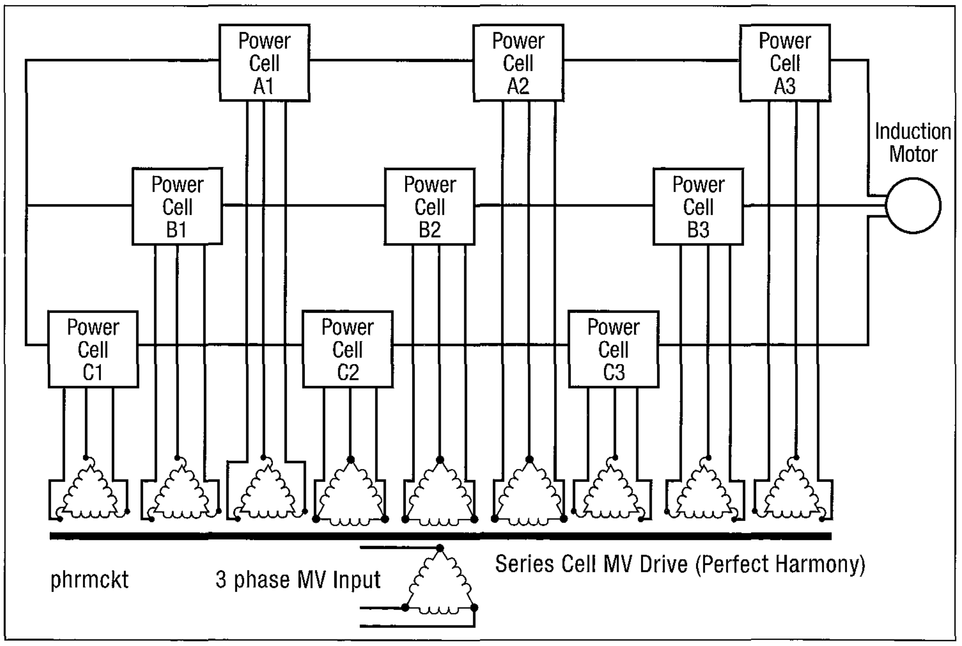

Figure 33. Series-cell multi-level VFD

MULTI-LEVEL SERIES CELL INVERTER

The patented series cell arrangement of Figures 33 and 34, also known as the Perfect Harmony drive, addresses the previously mentioned design issues in a unique way. Since there are no devices in series, only series cells, the problem of voltage sharing does not exist. The rectifier diodes and the IGBTs are both closely coupled to the DC link capacitor in the cell and thus cannot be exposed to more than the bus voltage, regardless of the load behavior. Since there is no DC link choke, a voltage transient on the AC mains is converted into a current pulse by the relatively high leakage reactance of the transformer secondary, and does not add to the voltage seen by the diodes.

Each cell generates the same AC output. The fundamentals are equal in magnitude and in phase, but the carrier frequency is staggered among the cells in a particular phase. See Figure 35 for the output waveforms. Although an individual cell operates at 600 Hz, the effective switching frequency is 3.6kHz, so the lowest harmonic is theoretically the 60th. This low switching frequency and the excellent high-frequency characteristics of the IGBT has the advantage that the IGBT switching losses are totally negligible. The devices can switch well above rated current without the need for snubbers which also helps in maintaining excellent efficiency. Waveform quality is unaffected by speed or load. For the 5 cell/phase VFD, there are ten 620 volt steps between the negative and positive peaks. With this technique, the

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concern for high dv/dt on the motor windings is avoided entirely.

A major advantage of the IGBT over all other power switches is the extremely low gate power required. The peak power is about 5 watts with an average of much less than 1 watt. This dramatically simplifies the delivery of gate power compared to the GTO/IGCT. Although there are more active devices in the Perfect Harmony (48 IGBTs and 72 diodes in the inverter sections) than in the other circuits, the elimination of snubbers, voltage sharing networks, and high-power gate drivers compensates for the additional switching devices. The type of IGBTs employed are third- and fourth-generation isolated base modules, generally the same mature product as those found in 460 VAC and 690VAC PWM drives, and are also used in traction applications. The IGBTs are protected by an out-of-saturation detector circuit which augments the built-in current limiting behavior. Since the cells are assembled into a non-conducting framework and are electrically floating, the mounting and cooling of the IGBTs is no more complex than in a low voltage PWM drive. It is possible to put redundant cells in the string, and also to operate at reduced output with one cell inoperative.

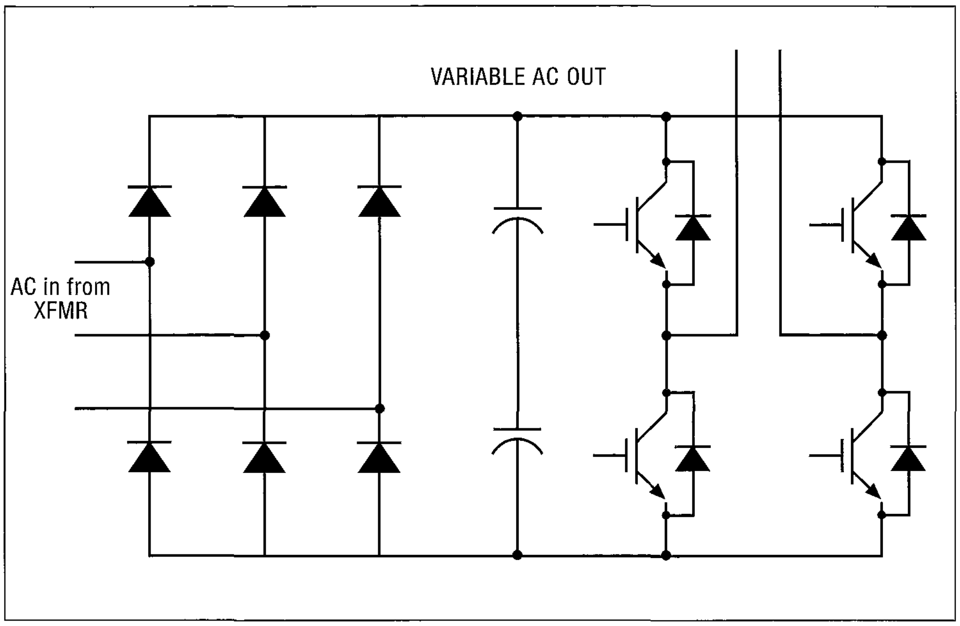

Figure 34. Power conversion cell for the series-cell multi-level VFD

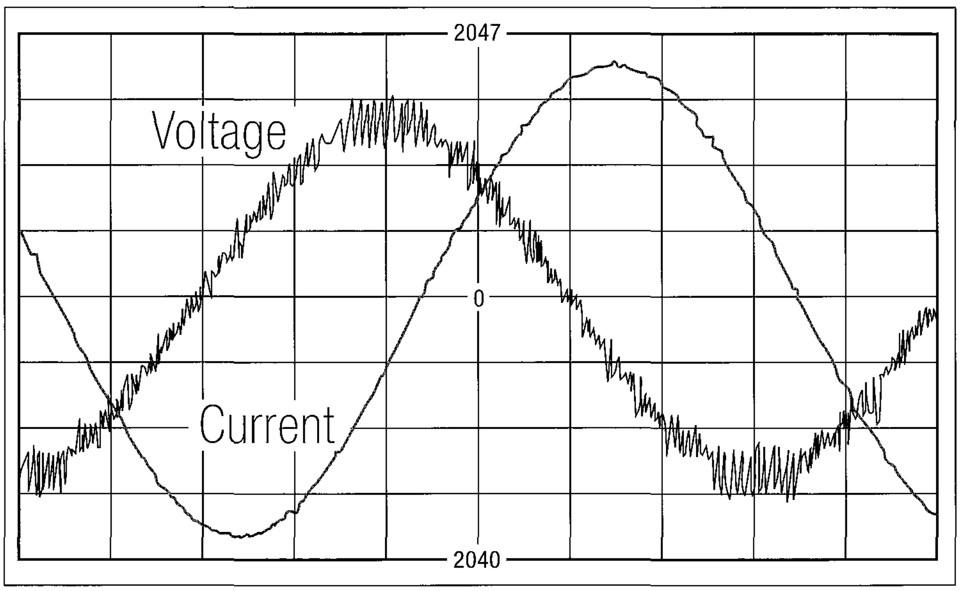

Figure 35. Output voltage and current of a series-cell multi-level VFD

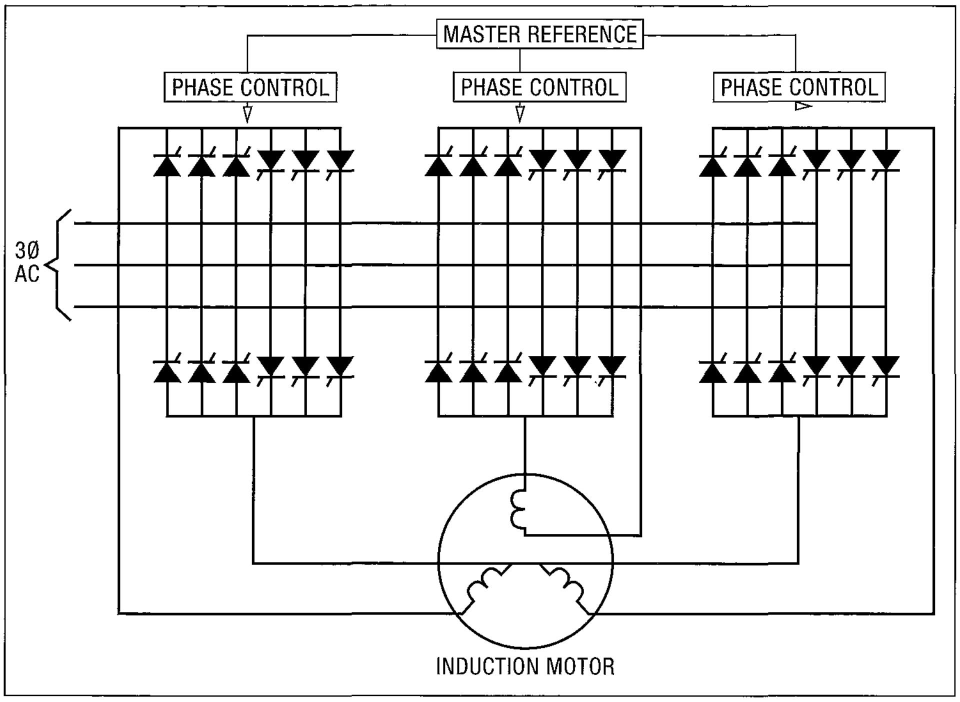

Figure 36. The 6-pulse cycloconverter motor drive

CYCLOCONVERTER

Still another approach in an IM drive is to "synthesize" an AC voltage waveform from small sections of the three-phase input voltages. This circuit arrangement is different from the previous types in that it does not have two conversion stages separated by a DC link. This requires at least three "dual converters," which are two thyristor bridges connected antiparallel, and the circuit is called a cycloconverter. See Figure 36. The output voltage is rich in harmonics but of sufficient quality for IM drives as long as the output frequency does not exceed 1/3 of the input frequency for a 6-pulse implementation. Twelve-pulse versions with more thyristors can generate better output waveforms at higher frequencies (closer to the line frequency). The thyristors are line commutated, but there are at least 36 of them. The cycloconverter is capable of extremely heavy overloads, fast response, and four-quadrant operation, but it has a limited output frequency and poor input power factor. The cycloconverter has been used very successfully for special low-speed, high-power (> 10MW) applications, such as cement-kiln drives and main rolling mill drives.

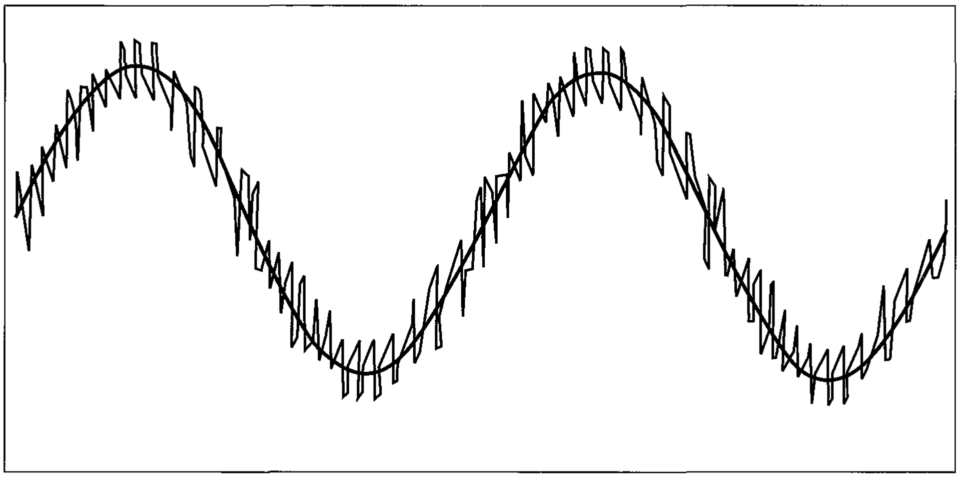

Figure 37. 25Hz output voltage wave and reference of 12-pulse cycloconverter (60Hz input)

COMPARISON OF MEDIUM VOLTAGE MOTOR DRIVES

All of the drive-types mentioned above are capable of providing highly reliable operation at a justifiable cost, and have been proven in service. They all have full load efficiencies above 95%. The most significant differences among them involve power quality; that is, how close the input current is to a sine wave, and how well the output resembles the sinusoidal utility voltage. Figure 38 compares a number of different factors. Voltage-fed drives have an advantage with regard to input harmonics and power factor, and the drives which do not use thyristors, have a wider speed range.

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	Load-commutated inverter	Neutral-point clamped inverter	Filter-commutated current-fed inverter	GTO/IGCT current-fed inverter	Multi-level series-cell inverter
Input harmonics	Fair(12-pulse) Poor (6-pulse)	Good (12-pulse) Very good (18pls)	Fair (12-pulse) Poor (6-pulse)	Fair (12-pulse) Poor (6-pulse)	Excellent
Uncorrected input power factor	Poor	Very good	Fair	Fair	Very good
Unfiltered output harmonics	Poor	Good	Good (near full speed)	Good	Excellent
Output common-mode V	High (fair)	None (excellent)	High (poor)	High (poor)	None (excellent)
Unfiltered output dv/dt	High (poor)	High (poor)	Low (good)	Low (good)	Low (good)
Regeneration capability	Yes	No	Yes	Yes	No
Torque pulsations	High (poor)	Low (very good)	Low (good)	Low (good)	Very low (excellent)
Special motor required?	Yes	No	Maybe (for common-mode V)	Maybe (for common-mode)	No
Speed range (PU)	.15 – 2.0	0 – 2.0	0.5 – 2	0 – 1.1	0 – 2.0
Special starting mode?	Yes	No	Yes	No	No

Figure 38. Medium voltage drive comparison

CONCLUSION

The improvement in process performance and energy savings is largely independent of the choice of drive. Although not always easy, the customer should compare a drive performance and price from a system perspective, as the particular drive choice has significant implications for the rest of the system, *e.g.*, on the motor and supply transformer.

A medium voltage drive customer has a wide choice with many feature options available. Undoubtedly, there will be improvements as new semiconductor switching devices are designed.

Biography of Richard H. Osman

Richard H. Osman received a BSEE degree from Carnegie Institute of Technology, Pittsburgh, PA, in 1965. He worked for Westinghouse Electric Corporation at the Research and Development Center from 1965 to 1970 where he was responsible for the development of a variety of solid-state variable speed drives, including thyristor DC drives, cyclo-converters and inverters.

Osman joined Robicon Corporation in 1970 as a Development Engineer in the DC drive group, where he designed special-purpose thyristor DC drives for earthmovers and transit vehicles. From 1977 to 1992, he was the Manager of AC Drives Engineering at Robicon. During this period his group developed a broad product line of both current-fed and voltage-fed type AC drives. From 1987 to 1988, Osman served as Technical Director of Heenan Drives Ltd., a sister company of Robicon located in Worcester, England. He also represented Robicon for five years on the NEMA Adjustable-Speed Drives Subcommittee and served as Chairman for two years.

From 1992 to 1994, Osman was Director of Drives Engineering at Halmar Robicon Group. From 1994 to 1996, he was Vice-President of Integrated Product Development, where he led Robicon in the development of the Perfect Harmony medium voltage drive.
From 1996 to 1998, Osman was Senior Vice-President of Technology for High Voltage Engineering, Robicon's parent company, where he served as technical advisor.

Today, Osman is ASIRobicon's Vice-President of Technology. He serves as technical advisor and works closely with the product development group.

Osman is a Senior Member of the IEEE, (The Institute of Electronic and Electrical Engineers) and is a member of the National Motors and Drives Steering Committee. Osman is a Registered Professional Engineer, who has written and presented more than 30 technical papers at various conferences and universities.

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